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06/07/02

PTO/SB/16 (10-01)
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PROVISIONAL APPLICATION FOR PATENT COVER SHEET

This is a request for filing a PROVISIONAL APPLICATION FOR PATENT under 37 CFR 1.53 (c).

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06/07/02

INVENTOR(S)

Given Name (first and middle [if any])	Family Name or Surname	Residence (City and either State or Foreign Country)
Michail	Tsatsanis	Santa Clara, California USA
Mark	Erickson	San Bruno, California USA
Ioannis	Kanellakopoulos	Cupertino, California USA

☒ *Additional inventors are being named on the* 1 *separately numbered sheets attached hereto*

TITLE OF THE INVENTION (280 characters max)

A METHOD AND SYSTEM FOR MULTILINE TRANSMISSION IN A COMMUNICATIONS SYSTEM

Direct all correspondence to: **CORRESPONDENCE ADDRESS**

☒ Customer Number [ ] → *08791*

OR *Type Customer Number here*

☒ Firm *or* Individual Name	Sanjeet K. Dutta Blakely, Sokoloff, Taylor & Zafman LLP				
Address	12400 Wilshire Boulevard, Seventh Floor				
Address					
City	Los Angeles	State	California	ZIP	90025-1030
Country	USA	Telephone	(408) 947-8200	Fax	(408) 947-8280

ENCLOSED APPLICATION PARTS *(check all that apply)*

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☐ Drawing(s) *Number of Sheets* ☐ Other (specify)

☐ Application Data Sheet. See 37 CFR 1.76

METHOD OF PAYMENT OF FILING FEES FOR THIS PROVISIONAL APPLICATION FOR PATENT *(check one)*

☐ Applicant claims small entity status. See 37 CFR 1.27.

☒ A check or money order is enclosed to cover the filing fees

☒ The Commissioner is hereby authorized to charge filing fees or credit any overpayment to Deposit Account Number: 02-2666

☐ Payment by credit card. Form PTO-2038 is attached.

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The invention was made by an agency of the United States Government or under a contract with an agency of the United States Government.

☒ No

☐ Yes, the name of the U.S. Government agency and the Government contract number are:

Respectfully submitted,

SIGNATURE Sanjeet Dutta

Date 06/07/2002

TYPED or PRINTED NAME Sanjeet K. Dutta

REGISTRATION NO. *(if appropriate)* 46,145

TELEPHONE (408) 947-8200

Docket Number: 1340P086Z

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INVENTOR(S)/APPLICANT(S)		
Given Name (first and middle [if any])	Family Name or Surname	Residence (City and either State or Foreign Country)
Thorkell	Gudmundsson	San Jose, California USA
Ming	Gu	San Jose, California USA
Sunil	Shah	Los Altos, California USA

Number 1 of 1

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PROVISIONAL PATENT

UNITED STATES PROVISIONAL PATENT APPLICATION

FOR

A METHOD AND SYSTEM FOR MULTILINE TRANSMISSION IN A COMMUNICATIONS SYSTEM

INVENTORS:
MICHAIL TSATSANIS
MARK ERICKSON
IOANNIS KANELLAKOPOULOS
THORKELL GUDMUNDSSON
MING GU
SUNIL SHAH

PREPARED BY:
BLAKELY, SOKOLOFF, TAYLOR & ZAFMAN LLP
12400 WILSHIRE BOULEVARD
SEVENTH FLOOR
LOS ANGELES, CA 90025-1026
(408) 720-8300

ATTORNEY DOCKET NO. 001340.P086Z

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Introduction

Copper infrastructure in the access network is ubiquitous and has served our voice communication needs well for over a century. As data connectivity needs have grown in the last twenty years, several technologies have been introduced to exploit this existing copper network and expand its usefulness. These include narrow-band modems, various versions of DSL, ISDN, DDS and T1/E1 technologies.

The Internet era has introduced new demands on the access network. The Gartner Group has estimated that relative to Metropolitan Area Network/Wide Area Network (MAN/WAN) needs, bandwidth demand is growing at 25% or more per year for the average enterprise.[1]

Figure 1 shows the relative growth of two traditional access services through 2006. A 25% growth curve is also shown for reference purposes.

Despite these projections, it is Voyan's belief that demand is constrained by the "service gap" – the void in services and pricing between T1 and T3. Creative services which address this gap in a cost effective way can increase penetration (total lines in service), margins and revenue.

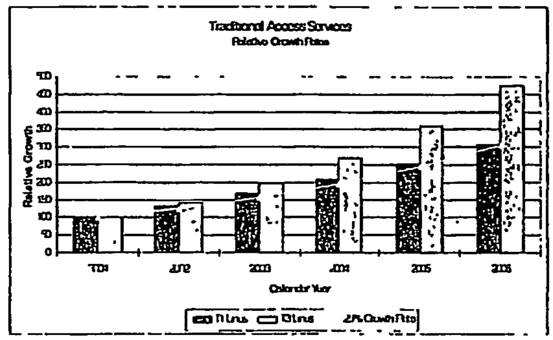


Figure 1

The Service Gap

T1/E1 service can now be delivered through a single copper pair using an HDSL2 or G.SHDSL modem. Data services at similar or higher speeds can be delivered using ADSL. Newer versions of this technology, such as VDSL, offer speeds of more than 50 Mb/s when operating in an asymmetrical mode (more bandwidth in one direction than the other). This is remarkable, given the fact that the copper plant was originally designed having only voice services in mind.

The difficulty has been in delivering high symmetrical speeds to all, or even the majority of customers, in a cost effective manner. For example, at 10 Mb/s or greater speeds, VDSL has significant reach limitations and can only serve a small percentage of the customer base. Before it can be widely adopted, a major and

[1] Gartner Group; *Look Out WAN – The Ethernet Roadkill Machine is Coming*; Research note COM-12-9201; Jay Pultz and Mark Fabb; February 6, 2001

costly re-engineering of the outside plant environment is needed to reduce the average loop length.

The vast majority of customers lie within a Customer Service Area (CSA) of 12,000 feet. At this range, symmetric 1.5 – 2.0 Mb/s service is close to the highest bit rate service that a single copper pair can deliver reliably. For speeds greater than this, fiber based services are most commonly deployed today.

Unfortunately, fiber is not ubiquitous in the access network. Current estimates are that less than 7% of all businesses can be reached by fiber. This is expected to increase to just over 10% by 2006.[2]

Construction of new fiber in the access network is typically focused on high density environments such as multi-tenant office buildings in large cities. But in the Internet era, the demand for high speed connectivity is widespread geographically.

Fiber is expensive to deploy. Construction costs can be significant as can the cost of the equipment itself. Organizations located "off net" often cannot justify the up-front construction costs of fiber, not to mention the dramatically higher monthly charges for fiber based service. For example, in 2001, total T3 costs per line were estimated to be 12 times higher than total T1 costs per line.[3]

This has created the "service gap". From an enterprise perspective, the jump from traditional copper to fiber based service is a large one. As customer demand for bandwidth continues to grow, this service gap will become increasingly apparent and problematic.

[2] Estimate based on Voyan research
[3] Frost & Sullivan; *US Broadband Services Access Market 2030-63*; 2000

The question is whether it is possible to provide these desired, high rate services using the existing copper infrastructure and close the services gap.

The answer is YES. Voyan has developed *OptiFusion™*, a technology which delivers optical performance levels by combining multiple copper pairs. It enables Carriers to offer these high rate services to the majority of their customers using the existing outside plant infrastructure.

OptiFusion™ technology uses multi-line techniques to dramatically improve the bandwidth that is achievable relative to traditional methods. An *OptiFusion™* enabled system can deliver speeds of 10 - 45 Mb/s over 4 – 10 copper pairs within a standard Carrier Service Area (CSA) under "real world" disturber conditions.

There are tremendous business benefits to the Carrier. High speed, copper based service can be delivered for less than half the cost of traditional T3 over fiber, significantly reducing capital expenditures and improving margins. True Fractional T3 (FT3) service and next generation Ethernet service is enabled and ideally suited for delivery using a multi-line technology platform. For the enterprise, Carrier services based on this technology can be available broadly and without high non-recurring costs.

The objective of this white paper is to explain the fundamentals of multi-line technology, present the performance gains that are achievable, and expand on the benefits that this delivers to a service provider. Specifically, the following topics will be covered:

- Limitations of copper
- Multi-line transmission
- Performance gains
- Carrier benefits

Limitations of Copper

Copper twisted pairs are usually of small gauge resulting in a significant signal reduction over long distances. Despite this attenuation, however, the capacity of a twisted pair at CSA range would be well above typical T1 rates if interference and noise could be suppressed. Unfortunately, copper pairs are typically not shielded and incur substantial ingress noise and interference from other lines. This is known as cross-talk.

Figure 2 illustrates the performance loss that is possible due to cross-talk interference. The red solid line shows the reach-rate curve (achievable rate versus loop length) for a modem complying with the SHDSL spectral mask, when no other services exist in the same binder. The blue dashed-dotted line shows the same reach-rate curve when one HDSL2 service is activated in the same binder. The degradation is evident resulting in a substantial bit rate loss.

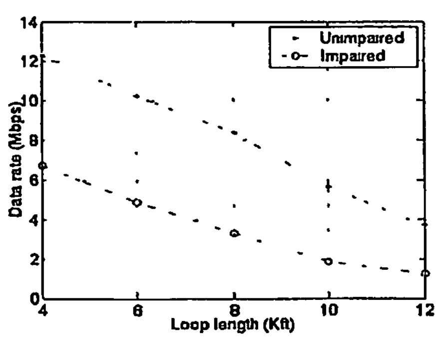


Figure 2

Similarly, ingress noise from RF sources, radio stations, electric motors etc. can result in significant performance degradation for the lines in an affected binder.

As will be explained, an advanced design of a multi-line transmission system can significantly mitigate those limiting factors and provide dramatic bit rate improvements.

Multi-line Transmission

The use of multiple copper pairs for higher data rates is not a new idea. The performance and cost effectiveness of products available today, however, is not sufficient for the Internet era. Too many pairs are required to achieve only a modest gain in performance. Therefore, these products are typically not used for Carrier network service offerings.

In contrast, the approach described here has to do with coordinated transmission of synchronized waveforms across multiple lines and the joint signal processing of all signals at the receiver. This Multiple Input Multiple Output (MIMO) technique takes advantage of the interdependencies of signals traveling across adjacent lines and results in impressive performance gains.

Cross-talk in a multi-line system takes two distinct forms, which require different technologies to address it. There is cross-talk originating from the system's own lines (in-domain cross-talk) and cross-talk originating from outside sources (out-of-domain cross-talk).

Figure 3 provides a pictorial view of this situation, where a number of lines in the binder belong to the multi-line system (red), while other lines belong to interfering services (purple).

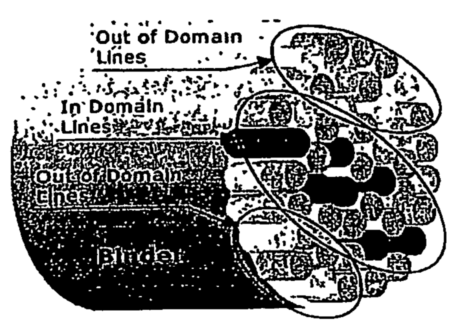


Figure 3

Near End Cross-Talk (NEXT) originating from in-domain lines is easier to cancel because the system has access to the interfering transmitter. Cancellation is based on identifying the precise filter that has to be applied to the interfering transmitted signal to match the exact opposite of the interference.

Figure 4 shows a simplified example with two lines, where NEXT cross-talk from Line 1 is cancelled at the receiver of Line 2. Notice that although the interference is introduced across the lines as the analog signals travel along the twisted pairs, the cancellation is performed in the digital domain through appropriate signal processing.

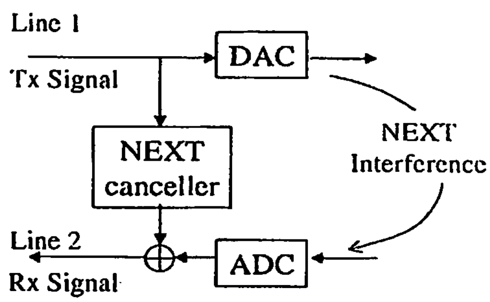


Figure 4

While Figure 4 illustrates the basic principle of in-domain cross-talk cancellation, it does not reveal all the complexities associated with a full matrix of cross-talk cancellers, from each possible transmitter to each possible receiver in a system with more than two lines. A number of innovative technical solutions have been developed at Voyan to make such an approach feasible within a reasonable computational complexity budget.

When cross-talk originates from out-of-domain lines, the receiver has no access to the interfering signal. Therefore, its mitigation is more difficult. Despite its difficulty, addressing out-of-domain interference is essential to achieving system performance. *Without it, in-domain cross-talk cancellation is of little use, because there is no guarantee that the in-domain interference component will always be stronger than the out-of-domain component.*

With *OptiFusion*™ technology, the mitigation of out-of-domain cross-talk and ingress noise is accomplished by the coordinated transmission across all lines. It involves joint MIMO processing of the signals across all lines both at the transmitter and the receiver. Furthermore, the solution includes joint Forward Error Correcting (FEC) coding of all signals across all lines.

Voyan treats the multi-line transmission medium as a MIMO channel and optimizes the transmitter and receiver processing for the given multi-line loop. In particular, it exploits the interactions of the transmitted signals across lines, as they travel along the loop. Furthermore, it exploits the interactions of the out-of-domain interference signals across lines and balances the receiver SNR in an optimal way.

This is achieved by trading off SNRs across lines and frequencies in order to reach the maximum data carrying capacity of the channel. This novel transceiver design is also able to mitigate ingress noise received across multiple in-domain lines.

OptiFusion™ includes state of the art multi-channel coding, which further improves the performance and protects the system from errors due to impulsive noise. The fact that signals from multiple

lines are jointly coded improves the robustness of the system against impulsive interference affecting one or more of the lines.

Multi-channel coding (also known as space-time coding) has received a lot of attention in the context of wireless multi-channel systems. Due to the random and unpredictable nature of wireless channels, multi-channel coding has proven to be an extremely valuable technology in improving the performance of wireless systems.

Compared to wireless links, copper based systems are fortunate in that the copper channel experiences variations at a much slower rate. This in turn allows reliable adaptation of both the receiver and the transmitter to the given channel characteristics as they slowly change. Therefore, it is possible to augment multi-channel coding techniques with channel-adapted, MIMO physical layer processing technologies.

Figure 5 illustrates the architecture of the Voyan multi-line transceiver and includes both the in-domain cancellation module and the transmitter and receiver MIMO processing and coding blocks.

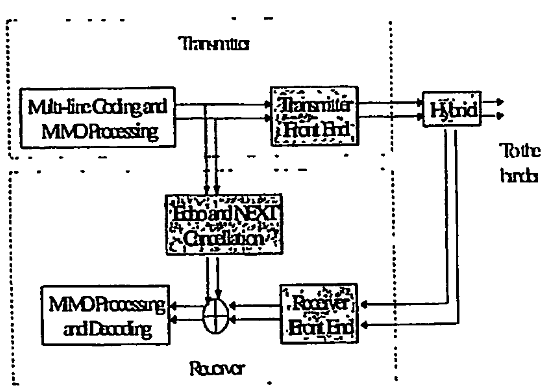


Figure 5

Performance Gains

The combination of MIMO processing, NEXT cancellation and multi-line coding results in impressive performance gains.

Figure 6 compares the SNR levels achievable with MIMO processing to those possible treating each line individually. The system in this example has 10 in-domain lines. Each line has length equal to 9,000 ft (26AWG) and each transmitter complies with the SHDSL spectral mask. Furthermore, the system is exposed to interference from 3 HDSL services.

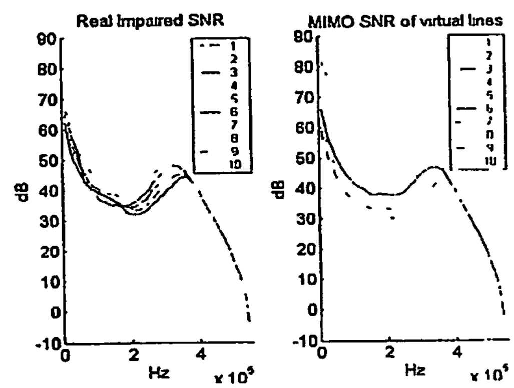


Figure 6

The plot on the left of figure shows the received SNR in each of the 10 lines as a function of frequency. The SNR degradation due to interference is evident in the midrange of frequencies. The plot on the right shows the improvements due to multi-line processing. The first four lines are now almost interference-free and the curves lie on top of each other. The rest of the lines experience only minor SNR changes.

The next figure illustrates how MIMO SNR gains translate into bit rate improvements. A system with 10 lines was simulated over 100 different random assignments of interference coupling transfer functions. With Figure 7, a more

severe interference environment is assumed consisting of 2 HDSL, 2 ADSL, 1 VDSL, 1 DDS and 2 HDSL2 services in the same binder.

The figure shows the resulting histogram of achievable data rates. Notice that the histogram is tightly clustered around 40 Mbps or an average 4 Mbps per line. In comparison, if the same 10 lines were used to provide T1 services, the aggregate data rate would have been only 15 Mbps as shown.

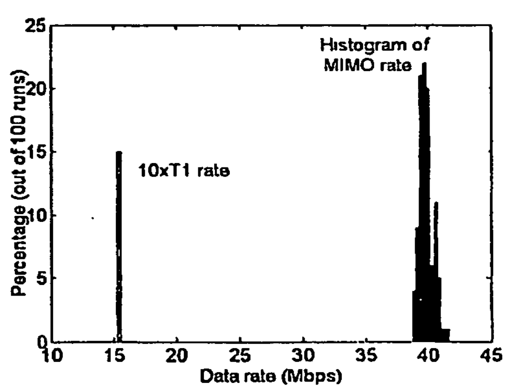


Figure 7

Carrier Benefits

Carriers will benefit from multi-line technology in many ways, ranging from improved financial performance to increased penetration. The exact benefits will depend on how the technology is utilized.

There are two broad strategies that enable a Carrier to derive benefit and there are synergies when both are employed.

The first strategy assumes no change to tariffs for T3, FT3 and related services.

The objective of the first approach is primarily financial gain - margin enhancement and cost reduction. Multi-line systems are expected to cost dramatically less than the SONET gear which is typically deployed today, reducing capital needs. Construction costs are also avoided, while service deployment can occur in a matter of days, speeding up time-to-revenue.

In this situation, a network planning organization would choose when to deploy multi-line technology instead of fiber. This would be done on a case by case basis, as customer orders are received.

For the Carrier, this represents an immediate benefit. Once a product is standardized and operations personnel are trained, it can be used right away. There are no tariffs to file or revise – and savings of 50% or more are immediately realized.

The second strategy offers all the same benefits, but can also increase revenue and penetration rates. The second strategy uses multi-line technology as a basis for creating a new set of differentiated access services.

For example, an Ethernet access service could be offered in all markets at speeds ranging from 10 Mb/s to 45 Mb/s. Pricing could be established in such a way that it did not cannibalize traditional services dramatically, but still created significant additional demand by filling the "service gap".

Similarly, a true fractional T3 service could be offered for customers who still desire an integrated voice/data solution. In addition to supporting standard private line access, the multi-line copper FT3 could be utilized by traditional packet services such as Frame Relay and ATM.

Attractive to Local Exchange Carriers of all types, these differentiated services introduce a "discontinuity" into the market

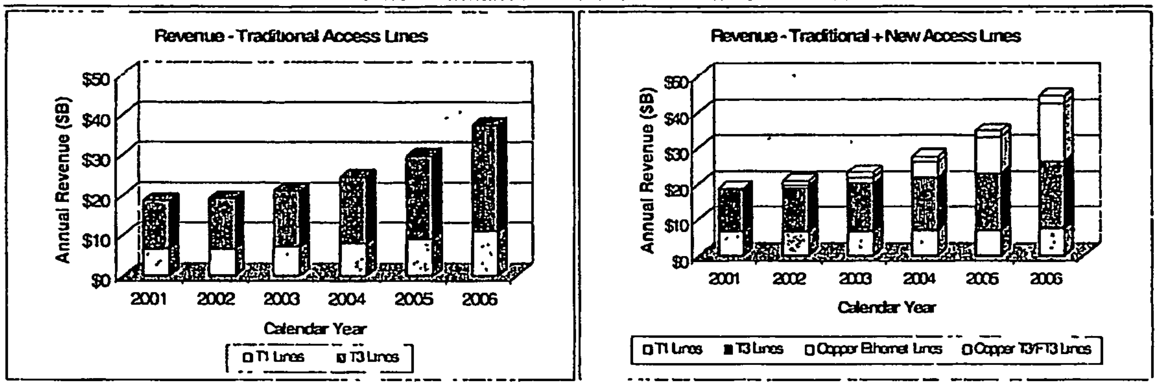
place and an opportunity to stimulate substantial new demand.

Figures 8 illustrates how penetration and revenues can be positively impacted by the introduction of new services using multi-line technology in the U.S. market. When combined with reduced deployment costs, these services enhance profitability and become an important part of the overall services mix.

Conclusions

Multi-line technology can fill the "service gap" that exists between traditional copper based and fiber based solutions. With *OptiFusion*™ technology, Carrier Class T3, Fractional T3 and "Ethernet First Mile" services can now be delivered without the need for fiber deployment. Increased revenue and improved margins will result from being able to deliver broadband services more quickly, more widely, and more cost effectively.

US Market - Enhanced Services Growth 2001 - 2006



US Market - Year End 2006 Summary Penetration and Revenue

Traditional Carrier Service Mix	*Penetration (Lines 000's)*	*Revenue ($B)*	*Traditional + Enhanced Carrier Service Mix*	*Penetration (Lines 000's)*	*Revenue ($B)*
Traditional T1	1,739	$ 11.0	Traditional T1	1,218	$ 7 7
Traditional T3	408	$ 26 6	Traditional T3	286	$ 18 6
			C- Ethernet Lines	772	$ 16 1
			C- T3/FT3 Lines	63	$ 2 1
Total	2,147	$ 37 6		2,339	$ 44.5
Multi-Line Technology Gain		***8.9%***	***Increased Penetration***		
Multi-Line Technology Gain		***18.4%***	***Increased Revenue***		

Figure 8

Multi-Pair and Multi-Access Capacity: Exploring the Potential

Motivation

- **Needs**
  - Capacity: 3-10 Mbps symmetric per pair
  - 3-6 pairs
  - Compatibility with legacy and basis systems-T1E1 DSM SG
  - Rate stability
  - CSA Range
  - Realistic impairments- e.g. PN4254
- **Multi-Pair and Multi-Access Capacities show potential for 2x-5x increase in rates**
  - Capacity computations provide bounds independent of implementation
  - Moore's law permits new approaches: 100x change in price-performance from 1995-2005
  - Research: Stanford, Princeton, industry, ...

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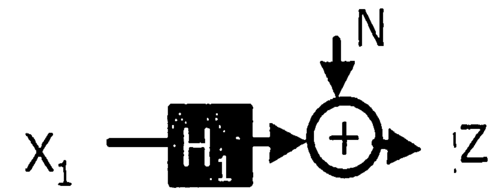
Different Formulations of Capacity Problems

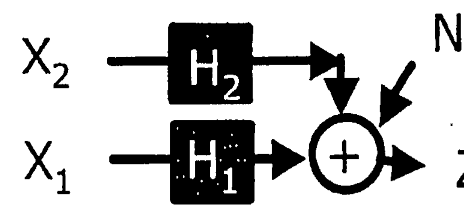
Gaussian disturber formulation (well researched)

- Transmit power constraint→ water-filling



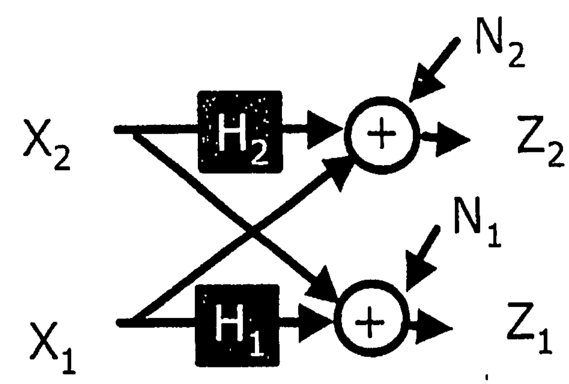
Multi-Access formulation (on-going research)

- Spectral and coding coordination, independent transmitters



Interference formulation (partial results)

- Green field versus Legacy



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Gaussian Channel Capacity

Single Pair (Scalar)

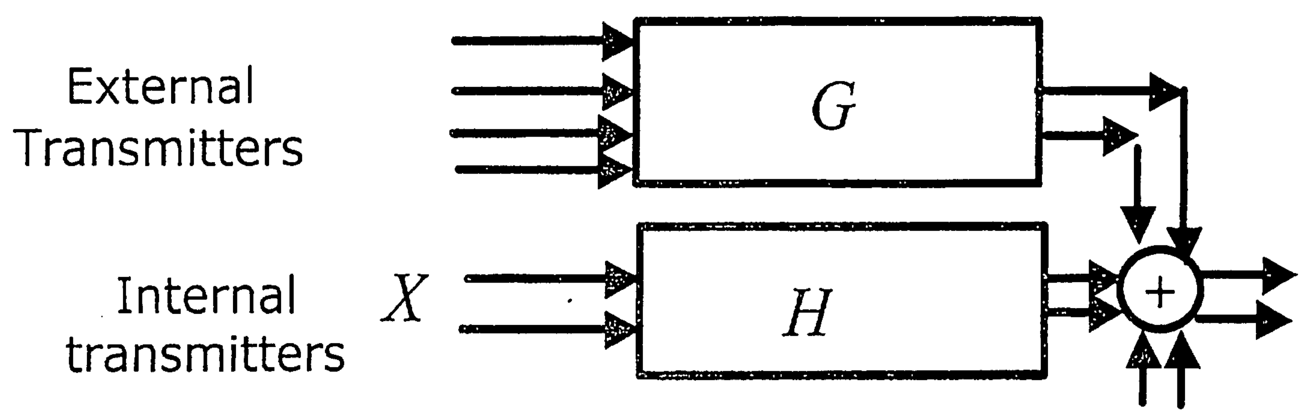
- *All* disturbers (including internal transmitters) in to a receiver are aggregated into a spectrally equivalent *scalar* Gaussian disturber

$$C=\sum_i \frac{1}{2}\int \log_2\left(1+\frac{\|H_{ii}\|^2 S_{xx_i}}{\sigma^2+\sum_{j\neq i}\|H_{ij}\|^2 S_{xx_j}}\right)df$$

Multi-Pair (Vector)

- *Internal* transmitters are vector inputs
- *External* disturbers are aggregated into a spectrally equivalent *vector* Gaussian process

$$C=\frac{1}{2}\int \log_2 \det(I+HS_{xx}H'R^{-1})df$$



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Gaussian Channel Capacities

Simulation Conditions

- **4-20 pairs at 9kft 26 AWG cable, *G.shdsl transmit mask***
- **AWGN –140 dBm/Hz**
- **Benefits of vector processing over scalar increase with Stronger Cross couplings**
  - Greater benefits with all couplings at Unger Mask
- ***Joint* distribution of N x N couplings becomes important**
- **Monte Carlo assignment of channels on *measured coupling matrices* from a 25 pair cable**
- **Maxima of measured couplings are at the Unger Mask**
- **Look at 99 percentile rates**

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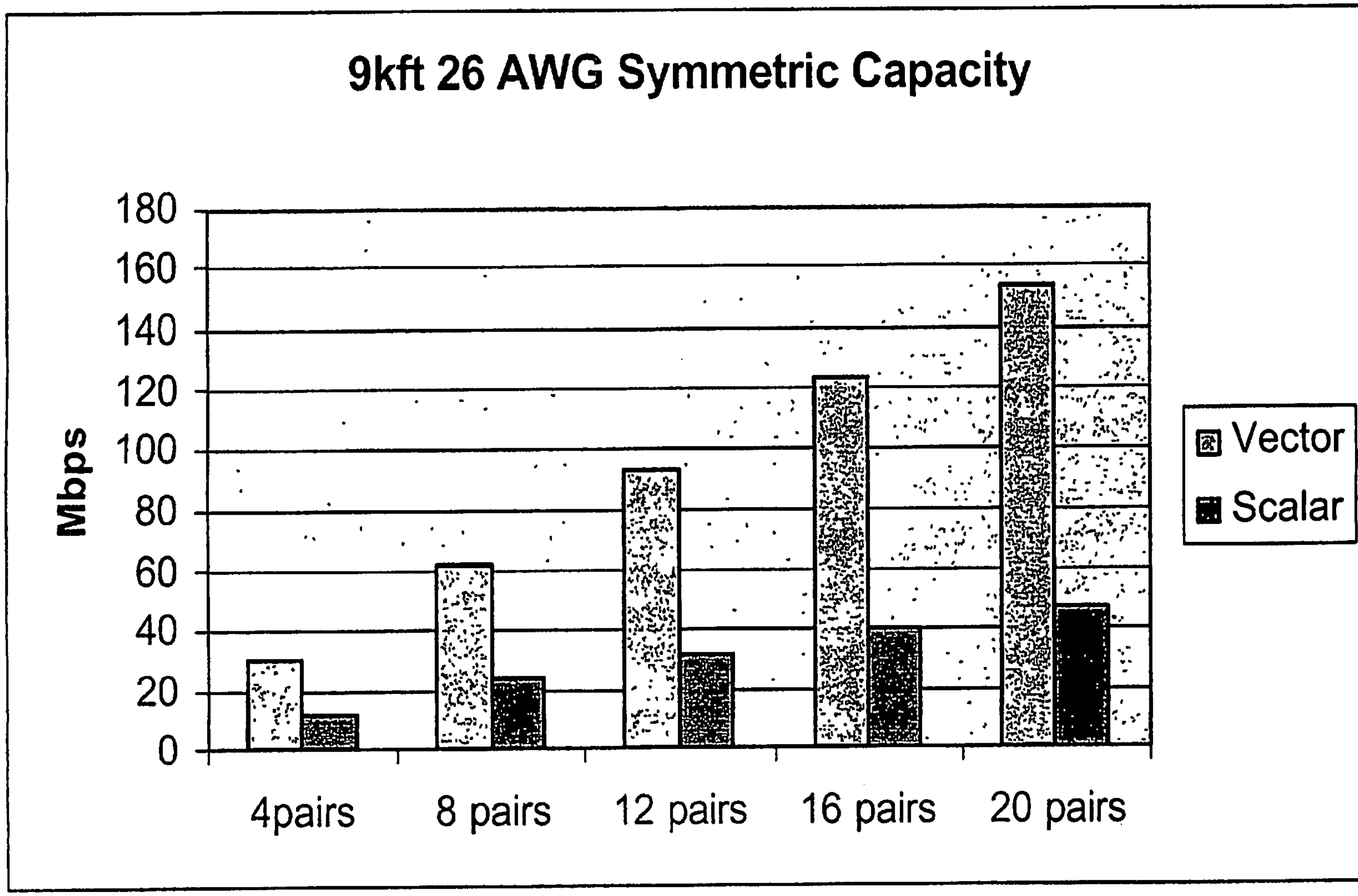
Typical Vector Capacity with Measured Cable Couplings



- Self FEXT and Self NEXT disturbance only, no external disturbers
- Monte Carlo results fall within 10% of typical

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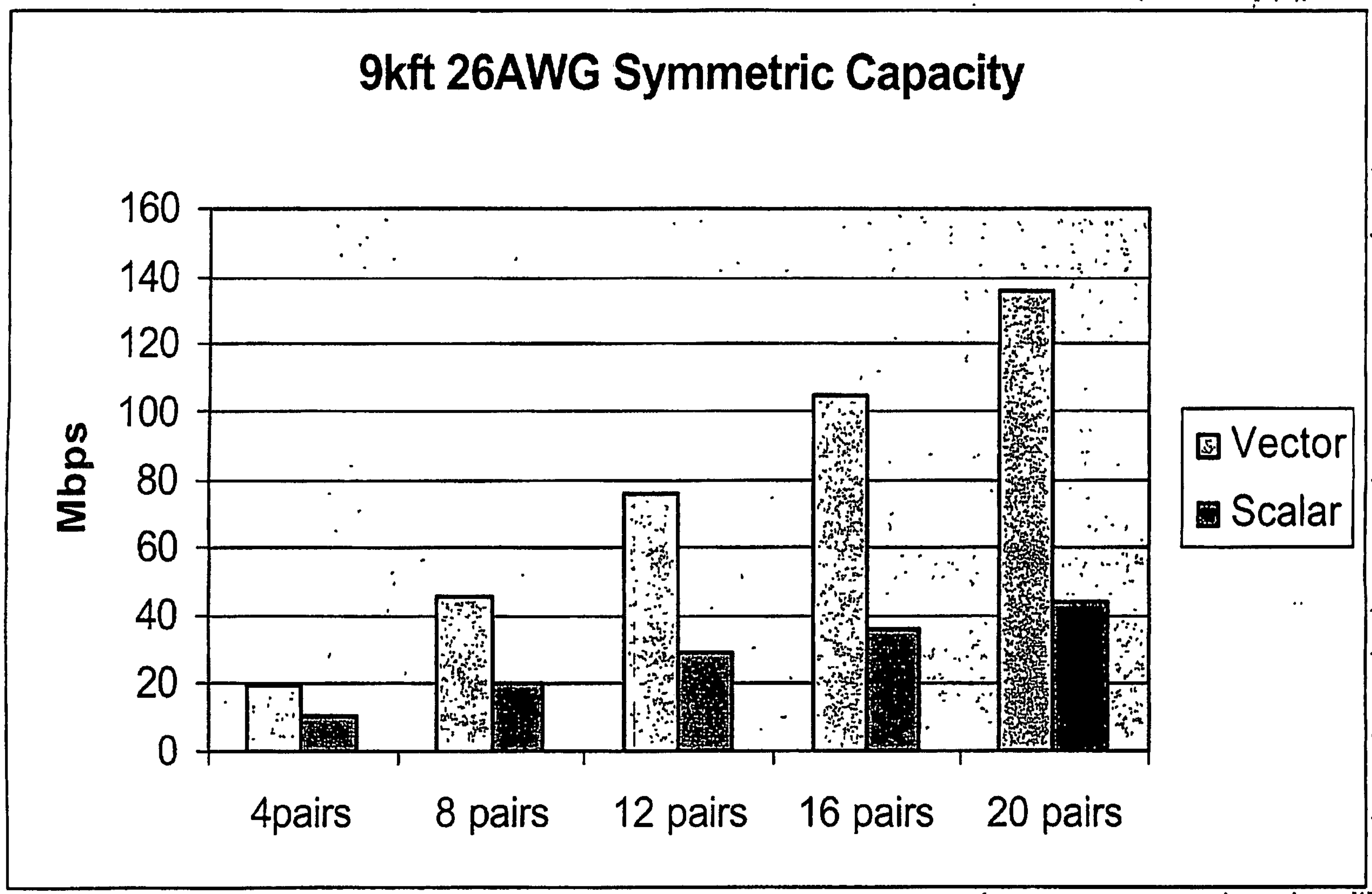
Typical Vector Capacity with Measured Cable Couplings



- Self FEXT, Self NEXT and 4 external disturbers (1 ADSL, 2 HDSL, HDSL2)
- Relative benefits increase with number of pairs

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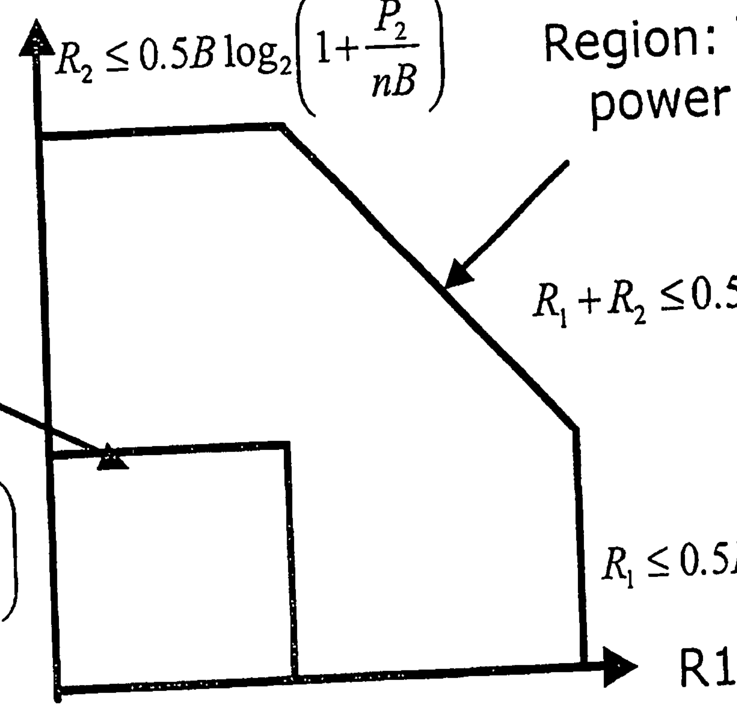
Multi-Access Formulation

Simplest Case (equal channel case)

Y = $X_1 + X_2 + N$

Achievable rate regions



$R_2 \leq 0.5B\log_2\left(1+\frac{P_2}{nB}\right)$

Multi-Access Rate Region: The other user's power adds to signal power!

$R_1 + R_2 \leq 0.5B\log_2\left(1+\frac{P_1+P_2}{nB}\right)$

$R_1 \leq 0.5B\log_2\left(1+\frac{P_1}{nB}\right)$

Single User Rate Region:

The other user's power adds to noise power

$R_2^* \leq 0.5B\log_2\left(1+\frac{P_2}{P_1+nB}\right)$

$R_1^* \leq 0.5B\log_2\left(1+\frac{P_1}{P_2+nB}\right)$

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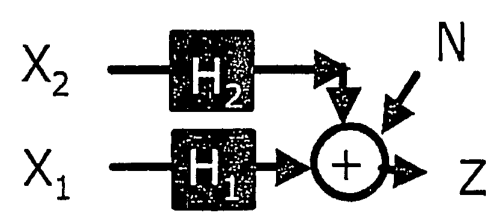
Multi-Access Formulation

General Multi-user case

- For unequal channels with ISI (*Cheng and Verdu, 1993*) showed rate regions with power constraints
- Iterative waterfilling converges to optimal for power constraints (*Yu, Rhee, Boyd, Cioffi, 2001*)
- *G.dmt.bis/G.lite.bis, T1E1.4/2001-200R5, T1E1.4/2001-278, T1E1.4/2001-284* address mechanisms to enable optimization



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Multi-User Detection

Multi-User detection exploits structure of disturbers

- Excess bandwidth
- Finite alphabet
- Receiver-only technology

Multi-User receivers can increase achievable rates beyond singleuser capacities in DSL

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Interference Problems

Interference limits

- Static capacity limits due to legacy and basis system
- Probabilistic future scenarios cover 95%-99% of cases
  - Useful for public-private combination - PN4254

Dynamic stability issues for further study

- Capacity calculations assume steady state
- Steady state requires convergence of adaptive schemes (e.g. bit swapping, TEQ/FEQ, DFE coefficients) in basis systems
- Convergence of adaptation needs to be ensured but is adversely affected by non-stationary noise
- Increased non-stationarity directly reduces the achievable rates compared to the steady state rates

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Summary

Benefits of Physical Layer Processing are in addition to those from higher layer aggregation or bonding

- Multi-Pair (vector) processing
- Multi-Access Methods

Future work needed to address issues

Very High-Rate Transport Over Multiple Copper Pairs

Problem Statement

Provide T3 rates over multiple copper lines at distances up to 8k – 12 k ft using a two-sided solution on a modest number of copper lines using Voyan's physical layer technology.

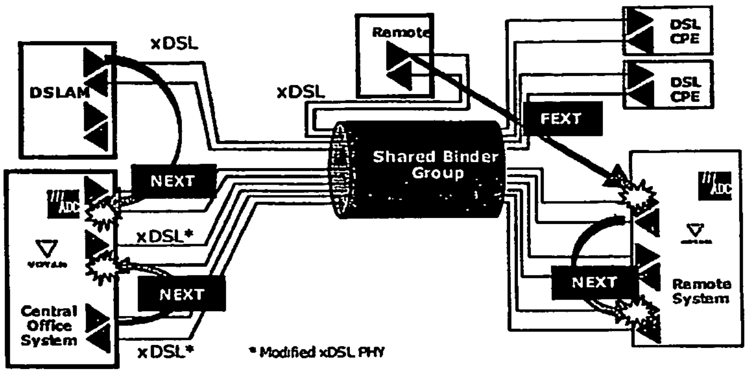


Figure: T3 Rate Transport of N-Pairs

The two sided solution need not be standards based; however, it must remain spectrally friendly to other services that are spectrally coupled to the pairs used by the service. To illustrate the limits of technology, this evaluation does not constrain computation resources.

Fundamental Limits and Regulations

Information theory provides a guide to fundamental limits of the copper pairs. Simply described, higher received power, higher transmit bandwidth and lower noise leads to higher capacity. However, as transmit power and bandwidth are increased, higher cross-talk is introduced in other services that share the binder. On-going spectral management standards activity defines mechanisms for ensuring reliability of services with certain confidence. In effect, the mechanisms rely on providing a *budget* for a certain amount of cross-talk spectral power into each type of service when deployed up to its allowed rate and reach.

Voyan Approach

Our proprietary approach is first to utilize the available spectrum more efficiently. This is accomplished using precise knowledge of cross couplings to other (foreign) services to reliably limit the actual ingress of spectral power into the other services to their allowable budget. These precise cross-couplings are used to optimize transmit spectra over multiple degrees of freedom simultaneously, enabling higher capacity without exceeding the allowable budget.

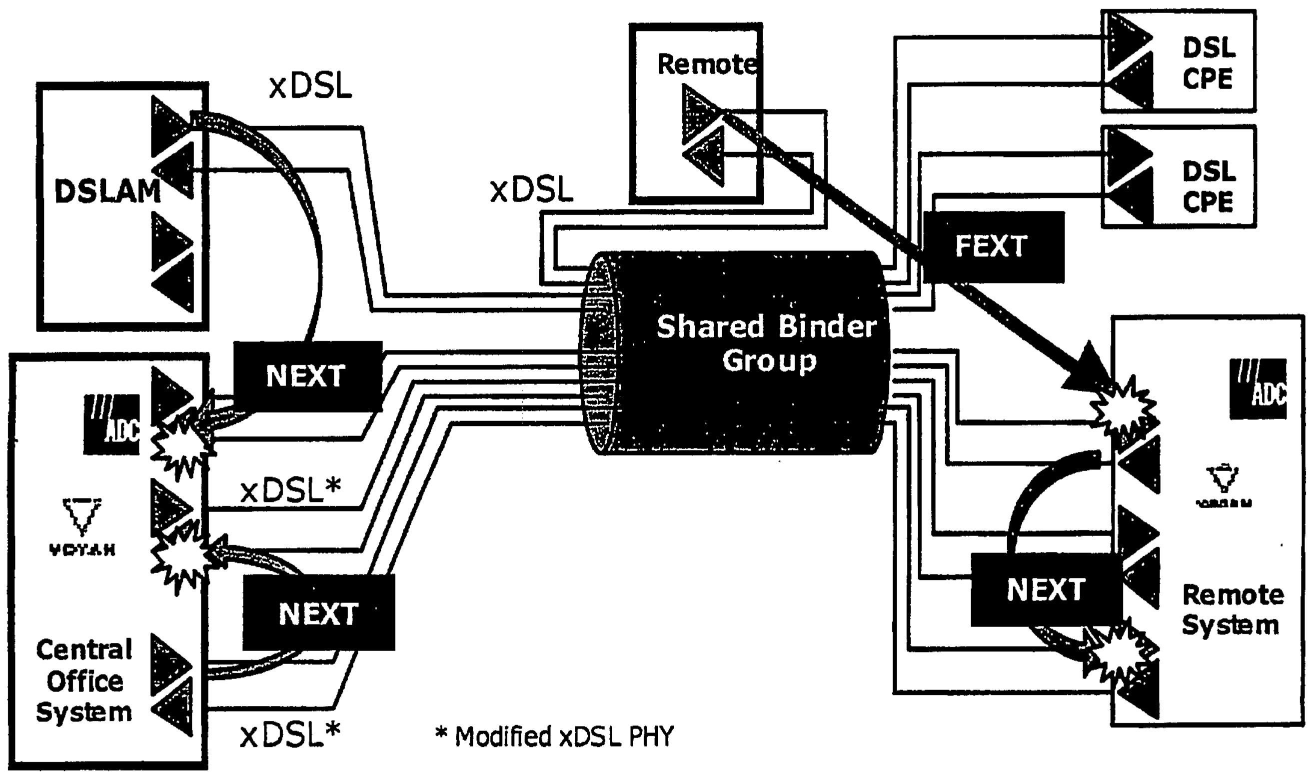
In addition, we use non-linear multi-user methods to significantly reduce the effective noise, including cross-talk from transmitters within our system as well as other (foreign) services. In doing so, we use cross-talk couplings accurately identified without requiring access to transmitted signals on foreign services. Multiuser methods exploit signal structures to separate cross-talk from the intended signals.

The following table provides representative achievable rates with our approach. Note that these rates do not constrain compute resources as described in the problem statement.

Number of Pairs	**Rate (Mbit/sec)**	**Loop Length (kft)**	**Gauge**
6	77	9	26
8	102.6	9	26
10	128.3	9	26
12	154	9	26
6	57.8	11.5	26
8	77	11.5	26
10	96	11.5	26
12	115.6	11.5	26

Very High Rate Transport Over Multiple Copper Pairs

Problem Statement



- T3 Rates over Multiple Copper Pairs at up to 8-12 kft
- Dual Sided Solution
- Technology Capability without Specific Compute Resource Constraints

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Results of Monte Carlo Analysis

6	77	9	26
8	102.6	9	26
10	128.3	9	26
12	154	9	26
6	57.8	11.5	26
8	77	11.5	26
10	96	11.5	26
12	115.6	11.5	26

Monte Carlo Analysis for Technology Capability

Main Channel

- EWL 9kft, 11.5kft no Bridge Taps for ease of comparison with baseline T1
- -140 dbm/Hz AWGN

Cross-talk Disturbance Scenarios

- 25 pair binder
- Measured Co-channels for NEXT
- Random Channel Assignments in Monte Carlo
- Self NEXT / FEXT
- Foreign NEXT / FEXT per TIA TR30.3 for business deployment

No Restrictions on Compute Resources

Spectral Ingress in to Foreign Services in the Binder Restricted to Current Standards

Basis for Higher Performance

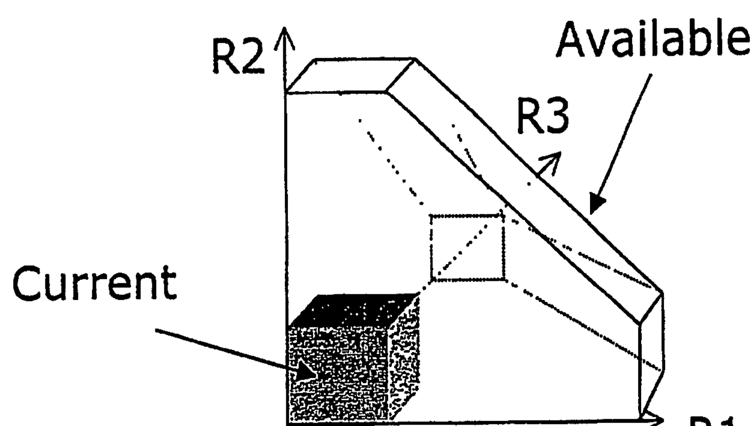
Current

Single User Capacity = B * $\log_2(1 + SNR)$

- **Higher power, Higher Bandwidth leads to Spectral Egress**
- **Cross-talk as Noise**

Available Capacity is Much Higher

- **Nonlinear Multi-User / MIMO Receiver and coding technology**
- **Exploit structure of out-of domain cross-talk and couplings to control egress**



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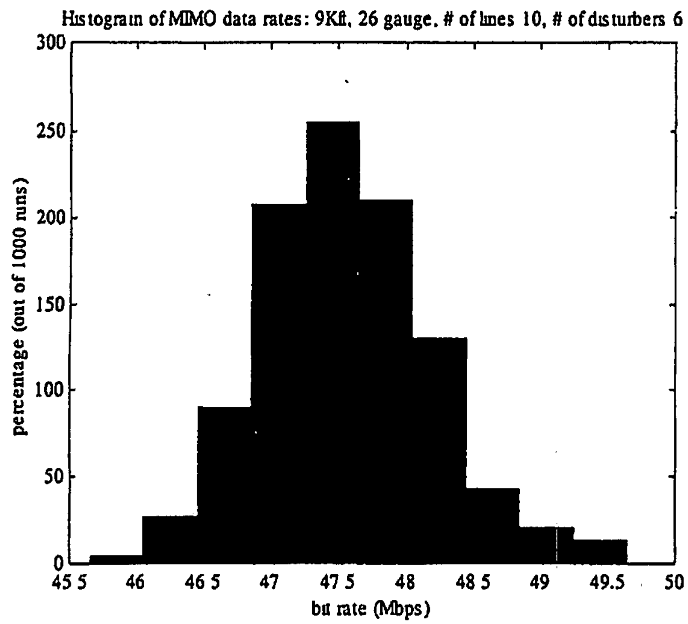
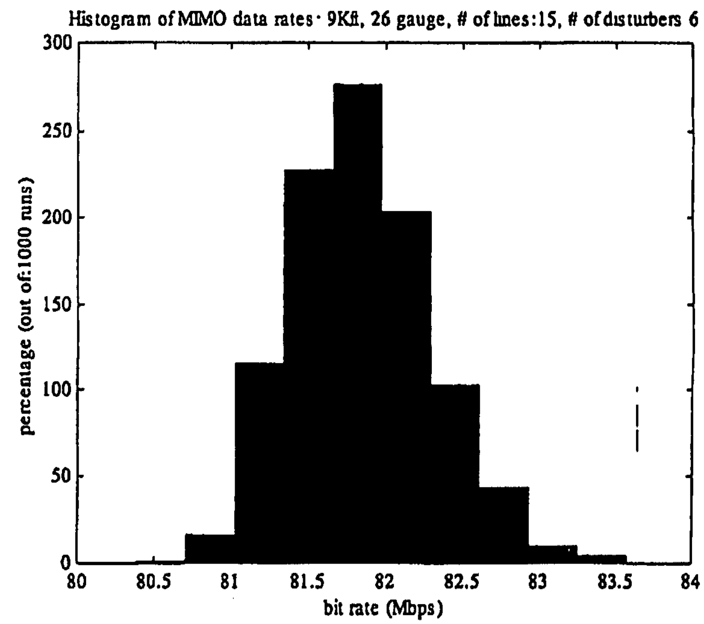
Reduced Complexity Example

Monte Carlo Analysis of rates:





Footprint Estimate: 135K Gates per line

Upside from Spectral Optimization can Increase Rates

First version with off-the-shelf components

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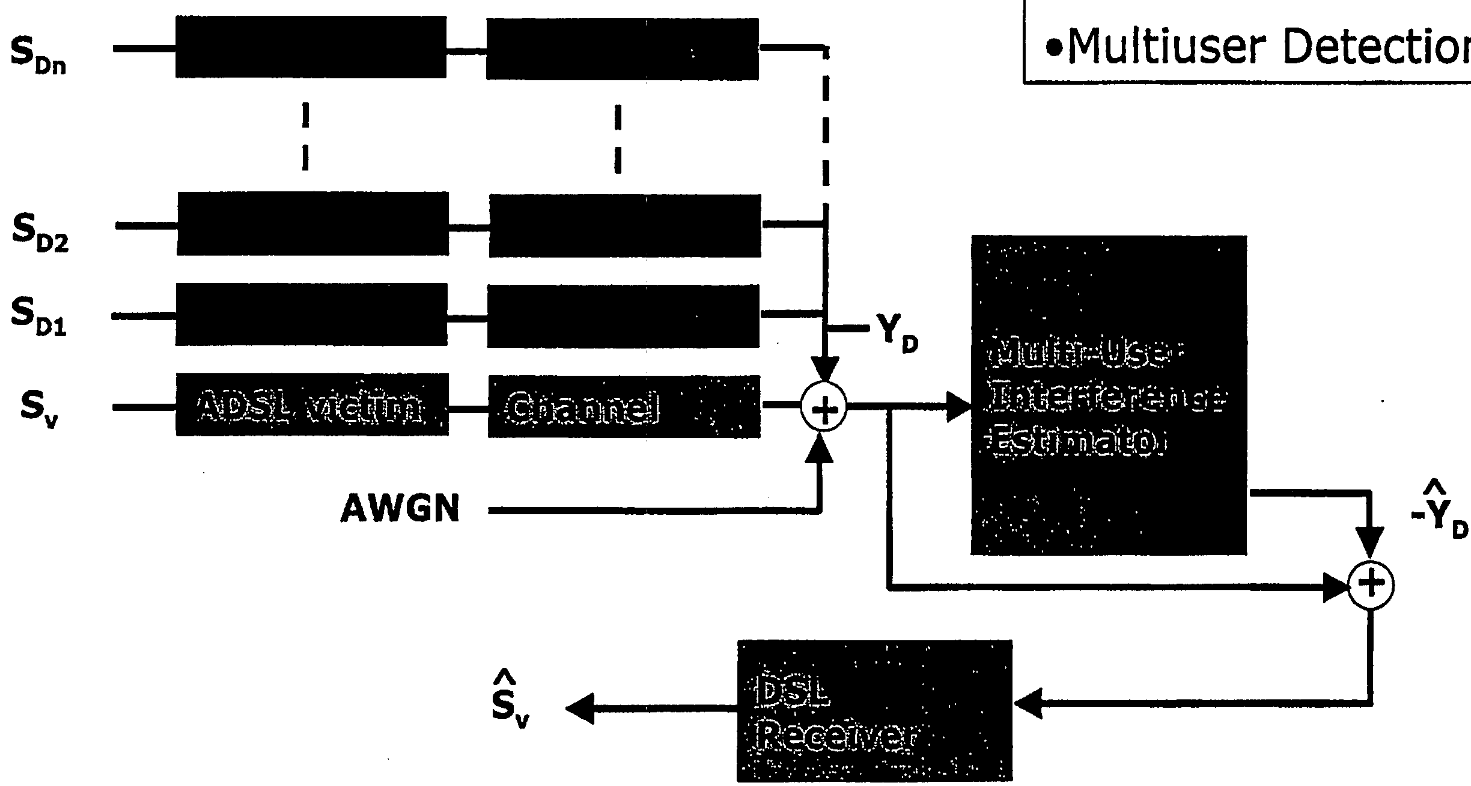
Crosstalk Compensation Example: ADSL CPE modem

Enabling Technologies:

- Co-channel identification
- Multiuser Detection



DSL Multiuser Channel

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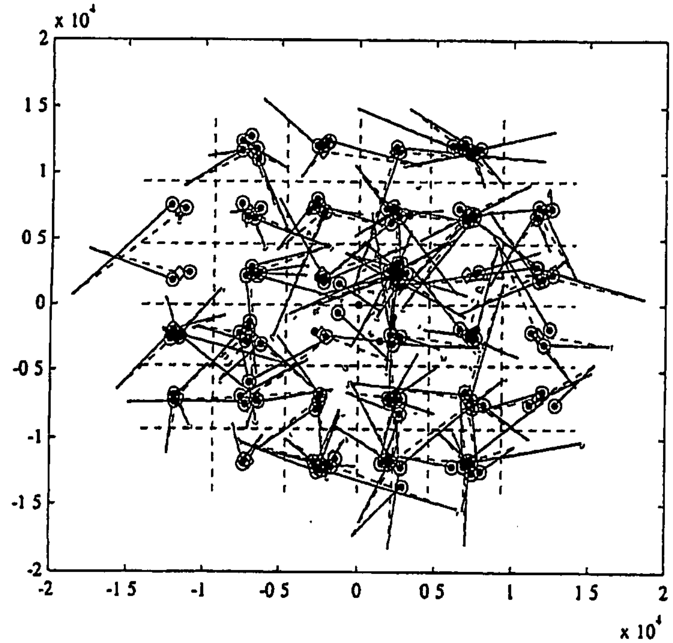
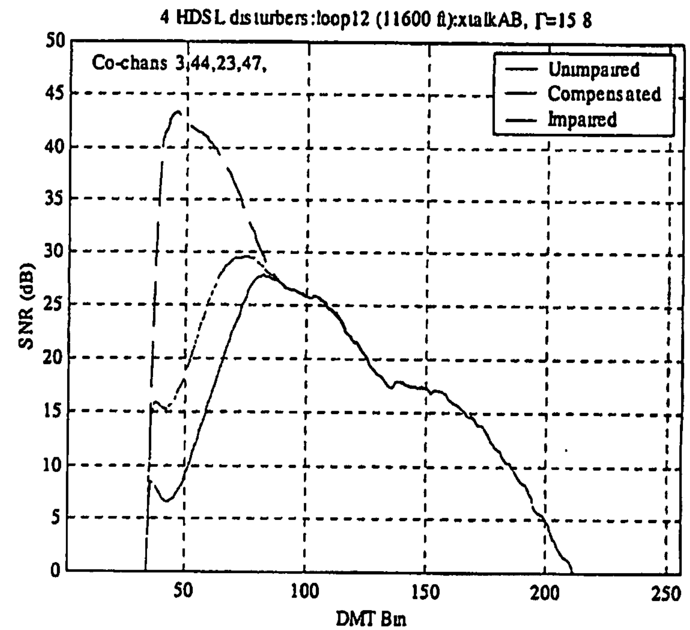
Crosstalk Compensation Example: ADSL CPE modem (cont.)

Pre- and Post-compensation Constellations & SNRs





More details: ITU contribution

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In the following description, for purposes of explanation, numerous specific details are set forth in order to provide a thorough understanding of the present invention. It will be evident, however, to one skilled in the art that the present invention may be practiced without these specific details. In some instances, well-known structures and devices are shown in block diagram form, rather than in detail, in order to avoid obscuring the present invention. These embodiments are described in sufficient detail to enable those skilled in the art to practice the invention, and it is to be understood that other embodiments may be utilized and that logical, mechanical, electrical and other changes may be made without departing from the scope of the present invention.

Some portions of the detailed descriptions that follow are presented in terms of algorithms and symbolic representations of operations on data bits within a computer memory. These algorithmic descriptions and representations are the means used by those skilled in the data processing arts to most effectively convey the substance of their work to others skilled in the art. An algorithm is here, and generally, conceived to be a self-consistent sequence of acts leading to a desired result. The acts are those requiring physical manipulations of physical quantities. Usually, though not necessarily, these quantities take the form of electrical or magnetic signals capable of being stored, transferred, combined, compared, and otherwise manipulated. It has proven convenient at times, principally for reasons of common usage, to refer to these signals as bits, values, elements, symbols, characters, terms, numbers, or the like.

It should be borne in mind, however, that all of these and similar terms are to be associated with the appropriate physical quantities and are merely convenient labels applied to these quantities. Unless specifically stated otherwise as apparent from the following discussion, it is appreciated that throughout the description, discussions utilizing terms such as "processing" or "computing" or "calculating" or "determining" or "displaying" or the like, refer

to the action and processes of a computer system, or similar electronic computing device, that manipulates and transforms data represented as physical (electronic) quantities within the computer system's registers and memories into other data similarly represented as physical quantities within the computer system memories or registers or other such information storage, transmission or display devices.

The present invention can be implemented by an apparatus for performing the operations herein. This apparatus may be specially constructed for the required purposes, or it may comprise a general purpose computer, selectively activated or reconfigured by a computer program stored in the computer. Such a computer program may be stored in a computer readable storage medium, such as, but not limited to, any type of disk including floppy disks, optical disks, CD-ROMs, and magnetic-optical disks, read-only memories (ROMs), random access memories (RAMs), EPROMs, EEPROMs, magnetic or optical cards, or any type of media suitable for storing electronic instructions, and each coupled to a computer system bus.

The algorithms and displays presented herein are not inherently related to any particular computer or other apparatus. Various general purpose systems may be used with programs in accordance with the teachings herein, or it may prove convenient to construct more specialized apparatus to perform the required method. For example, any of the methods according to the present invention can be implemented in hard-wired circuitry, by programming a general purpose processor or by any combination of hardware and software. One of skill in the art will immediately appreciate that the invention can be practiced with computer system configurations other than those described below, including hand-held devices, multiprocessor systems, microprocessor-based or programmable consumer electronics, network PCs, minicomputers, mainframe computers, and the like. The invention can also be practiced in distributed computing environments where tasks are performed by remote processing devices that are linked through a communications network. The

required structure for a variety of these systems will appear from the description below.

The methods of the invention may be implemented using computer software. If written in a programming language conforming to a recognized standard, sequences of instructions designed to implement the methods can be compiled for execution on a variety of hardware platforms and for interface to a variety of operating systems. In addition, the present invention is not described with reference to any particular programming language. It will be appreciated that a variety of programming languages may be used to implement the teachings of the invention as described herein. Furthermore, it is common in the art to speak of software, in one form or another (e.g., program, procedure, application...), as taking an action or causing a result. Such expressions are merely a shorthand way of saying that execution of the software by a computer causes the processor of the computer to perform an action or produce a result.

Disturber and Main Channel MIMO Processing

1. Executive Summary

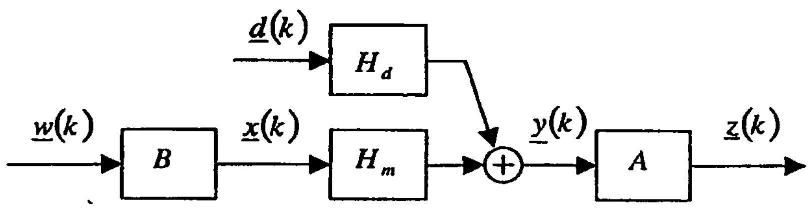
This document describes (in an incredibly brief fashion) a method for MIMO processing that satisfies two different criteria:

- Take care of in-domain FEXT (i.e., a non-diagonal main channel matrix)
- whiten disturber impairment across bins.

The system architecture is set up for one frequency bin across several pairs first. Next, the problem is stated. Finally, a solution is described.

2. System Structure

Block Diagram (in the Frequency Domain for One Bin)



Symbol	Definition
k	The kth DMT symbol
$\underline{w}(k)$	Vector of symbols for a single frequency bin
$\underline{x}(k)$	Vector to be transmitted in a single frequency bin after MIMO preprocessing
$\underline{y}(k)$	Received vector in a single frequency bin (before MIMO postprocessing
$\underline{z}(k)$	Vector of symbols to be decoded (after MIMO postprocessing) for a single frequency bin
$\underline{d}(k)$	Disturbance vector
H_m	Main channel matrix for a single frequency bin
H_d	Disturber matrix for a single frequency bin
B	Preprocessing Matrix
A	Postprocessing Matrix

Assumptions:

Prefixing works (e.g., no IBI).

H_m is a square full-rank matrix

The elements of $\underline{d}(k)$ are random variables with $E[\underline{d}(i)\underline{d}^H(j)] = R_d$ for $i = j$ and 0 otherwise.

3. Problem Statement and Solution

Goal: Compute a pair of matrices A and B with the following properties:

1. B is Hermitian (power is preserved across pairs by preprocessing)
2. $AH_mB = I$ (the received signal can be sliced properly)
3. $E[AH_d\underline{d}(i)\underline{d}^H(j)H_d^HA^H]$ is diagonal for $i = j$ and 0 otherwise.

Solution:

1. Compute C such that $CH_m = I$.
2. Compute B such that $B^HCH_dR_dH_d^HC^HB$ is diagonal and $B^HB = I$.
3. Set $A = B^HC$

Notes:

- Goal 1 is clearly satisfied by Solution Step 2, which can always be done.
- Goal 2 can be verified by substituting the equalities in Solution Steps 1 and 3 into the expression in the goal: $AH_mB = B^HCH_mB = B^HB = I$.
- Goal 3 is clearly satisfied by Solution Steps 2 and 3.
- The Matrix C can be interpreted as an FEQ that is MIMO across pairs, but not across bins. When there is no self-FEXT (i.e., the matrix H_m is diagonal), C can be implemented via a collection of SISO FEQ filters, one per channel.

A Reduced-Complexity Solution for the Fat Pipe System

This document presents a technical description of a proposed solution for the fat pipe system. It is focused on a computationally simple architecture, which should provide reasonable gains without the need for full-blown PAM receivers and their associated complexity.

Problem Statement

Let us assume that a vectored DMT system is in place utilizing M lines with appropriately trained TEQs, so that each line's channel is diagonalized across frequency bins. Let us also assume that those lines are impaired by K "out of domain" disturbers and each of those disturbers may impair multiple lines (for the purposes of this discussion, "in domain" impairment is considered to be perfectly removed through the use of NEXT echo cancellation techniques). Then, the received vector signal

$\boldsymbol{y}(\omega) = [y_1(\omega), \ldots y_M(\omega)]^T$ on those M lines and the particular frequency bin ω is given by

$$\boldsymbol{y}(\omega) = \boldsymbol{D}(\omega)\, \boldsymbol{x}_{dmt}(\omega) + \boldsymbol{v}(\omega)$$

where $\boldsymbol{x}_{dmt}(\omega)$ is the transmitted DMT signal, $\boldsymbol{D}(\omega)$ is the MIMO channel response and $\boldsymbol{v}(\omega)$ is the interference-plus-noise term consisting of

$$\boldsymbol{v}(\omega) = \sum_{k=1}^{K} \boldsymbol{h}_k(\omega)\, s_k(\omega T) + \boldsymbol{u}(\omega)$$

Here $\boldsymbol{h}_k(\omega)$ is the vector signature of that disturber across the M lines, $s_k(\omega T)$ is the discrete time Fourier transform of the disturber transmitted sequence, T is the Baud rate and $\boldsymbol{u}(\omega)$ is additive white Gaussian noise. All disturbers are assumed to be of the same Baud rate in this discussion.

A few remarks on this model are now in order:

- At the target distances of 10Kft, FEXT should be negligible and therefore the main channel MIMO response $\boldsymbol{D}(\omega)$ should be diagonal. Our analysis however is not limited to diagonal channels, and could handle short loops with significant FEXT. We will allow then $\boldsymbol{D}(\omega)$ to be a full matrix here.
- The DMT signal model is exact but the disturber impairment model is only approximate. The reason is that the disturbers do not have a prefix. To be exact, one more term should be added representing the "edge effects" where PAM pulses only partially overlap with the current DMT symbol.
- As a matter of notational convenience, we will drop the explicit dependence on frequency ω in what follows.

Optimal Signaling on the MIMO Channel

Let us investigate what should be the optimal modulation and signaling strategy for the interference environment at hand. For reasons of reduced complexity, we would like to limit the scope of this investigation to solutions, which do not explicitly utilize the finite alphabet property of the disturbers. In that sense they are in fact sub-optimal. They would be optimal only in the case the disturber transmitted signal were Gaussian. The proposed solution, although sub-optimal, obviates the need for developing SHDSL compensation architectures and can also handle ingress and AM noise.

One could argue that if the channel matrix is diagonal, then the problem can be decomposed into M independent problems to be solved separately. Then, the per-line water-filling bit-loading algorithm in the DMT standard is nearly optimal and there is nothing more one could hope for. The fallacy of this argument lies in the fact that the interference-plus-noise term is not white and hence introduces dependencies among the lines. In particular, the correlation matrix of that term is

$$R_v = E\{ v v^H \} = \sum_{k=1}^{K} h_k h_k^H + \sigma_u^2 I$$

Let us then study the equivalent problem with a pre-whitened signal. Let us consider the eigendecomposition of R_v, written as $R_v = U_v \Lambda_v U_v^H$ with obvious notation, and let the receiver pre-whiten with the matrix $\Lambda^{-1/2} U_v^H$. Then, the whitened signal is

$$y_w = \Lambda^{-1/2} U_v^H D x_{dmt} + v_w$$

where v_w is the whitened noise with $R_{v_w} = I$. Notice that the whitened signal model above involves an equivalent DMT MIMO channel equal to $H_{eq} = \Lambda^{-1/2} U_v^H D$. Let us denote the SVD of that channel as $H_{eq} = U \Sigma V^H$. Then, information theoretic principles indicate that optimal signaling across this channel involves prefiltering at the transmitter by V and postfiltering at the receiver by U^H. Substituting $\overline{y} = U^H y_w$ and $x_{dmt} = V \overline{x}_{dmt}$ into the whitened signal model we obtain

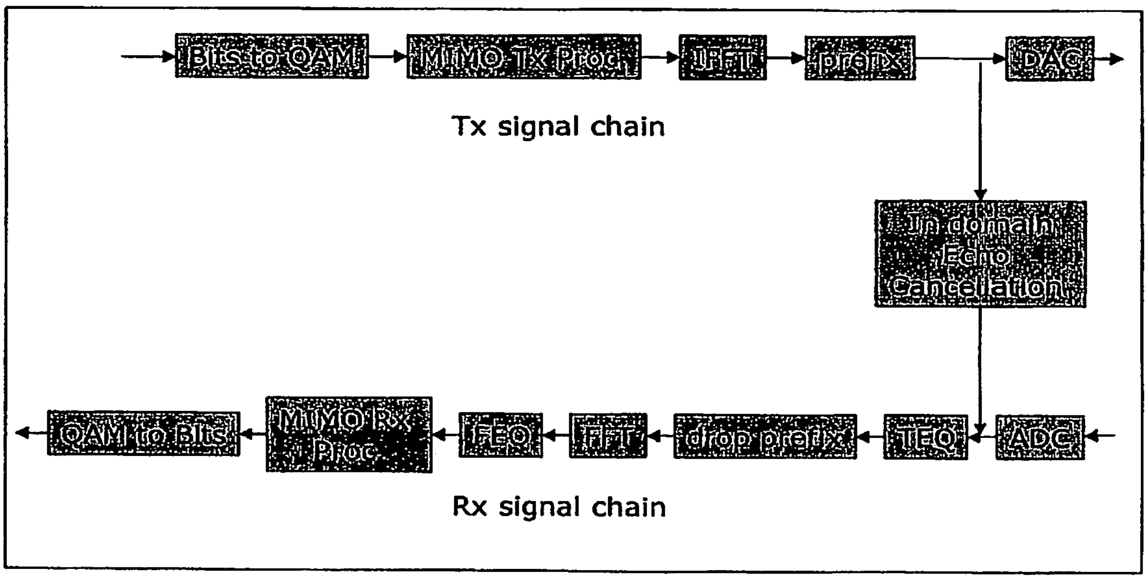
$$\overline{y} = \Sigma \overline{x} + \overline{v}$$

where $\overline{v} = U^H v_w$ is white. Therefore, the optimal signaling scheme in this environment (given a total power budget constraint) is water-filling bit-loading across the singular values on the diagonal of Σ.

In conclusion, if the above channel diagonalization procedure is applied to each vector bin signal, the system's capacity and/or achievable bit rates for a given SNR gap can be calculated through a two dimensional waterfilling procedure across frequency bins and copper lines.

Optifusion Block diagram

IFFT: Inverse Fast Fourier Transform
DAC: Digital to Analog Converter
ADC: Analog to Digital Converter
TEQ: Time Equalizer
FFT: Fast Fourier Transform
FEQ: Frequency-domain Equalizer
MIMO: Multi-Input Multi-Output
QAM: Quadrature Amplitude Modulation



Transceiver

Title: Optimal Receiver Window Shaping for Multiline Systems

Background: FFT-based Discrete Multitone (DMT) transceivers suffer from the "edge-effect" problem. In single-pair systems, this problem manifests itself as "spectral noise leakage" from bins where the noise power is high to bins where the noise power is low. A typical example of this problem is the deterioration of DMT performance in the presence of AM radio-frequency disturbers: even though their spectrum is narrow and should thus affect only 2-3 bins, AM disturber noise "leaks" into many more bins and becomes a very serious problem.

These "edge effects" are even more detrimental in the case of the OptiFusion multi-pair system, because, in addition to spectral leakage, they cause the disturber noise power to spread across multiple virtual lines, thus reducing the effectiveness of the MIMO processing and thus the overall data rate of the multi-pair system.

Edge effects are caused by the fact that the FFT that converts the received samples from the time domain to the frequency domain is of finite length: To demodulate a received DMT symbol, a time-domain signal equal in length to the symbol must be processed with an FFT. The shorter the DMT symbol, the shorter the FFT. The rate at which data can be transmitted through a DMT system is dependent on the Signal to Noise Ratio (SNR) at each frequency. The finite length of the FFT is equivalent to convolving the spectrum of the noise with a Bartlett window, which "smears" the noise power across many frequency bins. This is not a problem when the noise spectrum is flat, for example with white noise. But when the noise spectrum varies greatly with frequency, as is the case with narrowband AM disturbers, this "smearing" greatly reduces the SNR of bins where the noise would have been very low if the FFT had infinite length. The shorter the FFT, the greater the noise power amplification in these clean bins, and the greater the reduction of SNR and, thus, total bitrate.

Description: The invention proposed here is a time-domain processing stage that reduces the impairment caused by the finite-length FFT in a single-pair or multi-pair system. In simple terms, it combines the last few samples of data in the prefix (just before the DMT symbol begins) with the last few samples of the DMT symbol. The way the data is combined is defined through a "shaping function." The shaping function is constructed to achieve two goals:

1. the transmitted signal is reconstructed perfectly; and

2. the "smearing" of the noise power (as it appears at the output of the FFT) is minimized.

The invention also includes an optimization method for computing a shaping function that maximizes the bitrate of a multi-pair system.

Optimal Receiver Window Shaping

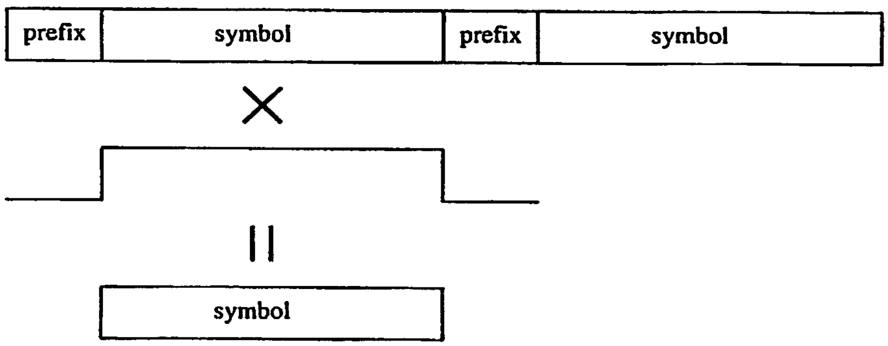
1. Executive Summary

In this document, we:

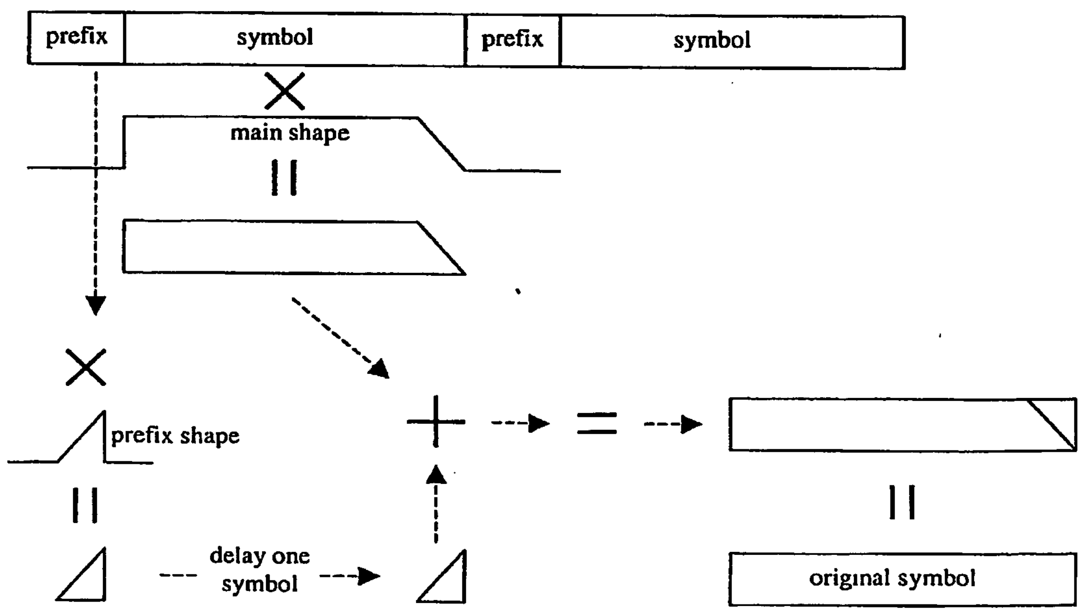
- describe the receiver window shaping structure in simple graphical terms.
- pose an optimization problem for finding the best window
- show numerical results of the optimization

2. Receiver Window Shaping

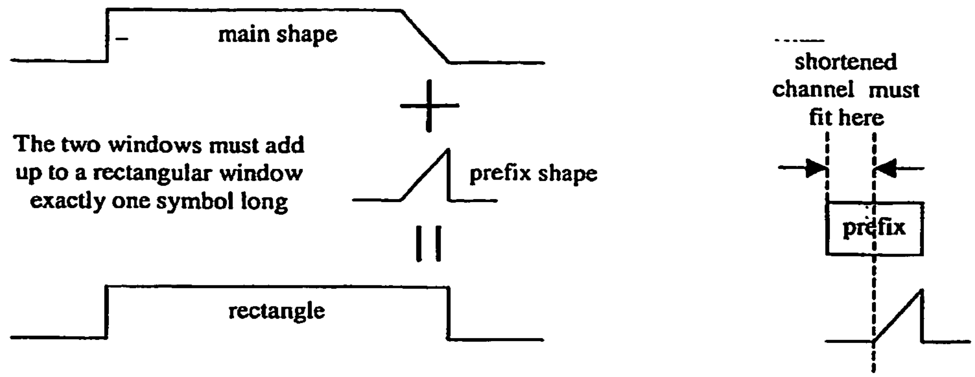
"standard" ADSL Window



Receiver Window Shaping



Key Insight: Perfect reconstruction of the signal sent with a prefix is assured if:



3. Finding an Optimal Shape

From the graphical analysis above, we are free to change the prefix shape and main shape as long as we satisfy the following two conditions:

1. The prefix shape and main shape add up to a rectangle
2. The length of the equalized main channel and the length of the prefix shape added together are less than or equal to the length of the prefix.

Under these conditions, the signal strength is invariant with respect to the prefix shape. Since the noise does not have a prefix, however, it is not invariant with respect to the prefix. To formulate a relevant optimization criterion, we first define the noise covariance matrix for the *i*th bin of a multichannel system to be R^i. The *j*th singular value of this matrix is σ_j^i. For a multiline system with MIMO processing, the SNR in dB of the *i*th bin of the *j*th virtual line is given by the expression

$$SNR_j^i = 10\log_{10}\left(p^i\right) - 10\log_{10}\left(\sigma_j^i\right) \qquad (1)$$

where p^i is the power transmitted in the *i*th bin of the virtual line. This power is fixed by selecting a spectral template consistent with spectral compatibility standards.

The number of bits that can be transmitted in *i*th bin of the *j*th virtual line is proportional to the SNR in dB. To maximize the overall bitrate for the multiline system, we would like to maximize the sum of the SNR's over all bins and all virtual lines. Since the powers p^i for the bins are fixed, we can maximize the overall bitrate by minimizing the double sum:

$$\sum_{j=1}^{N_{lines}} \sum_{i=1}^{N_{bins}} 10\log_{10}\left(\sigma_j^i\right) \qquad (2)$$

4. Experimental Results

To obtain experimental results, we constructed a function that computes the covariance matrices for a single PAM disturber. The function takes as inputs the following:

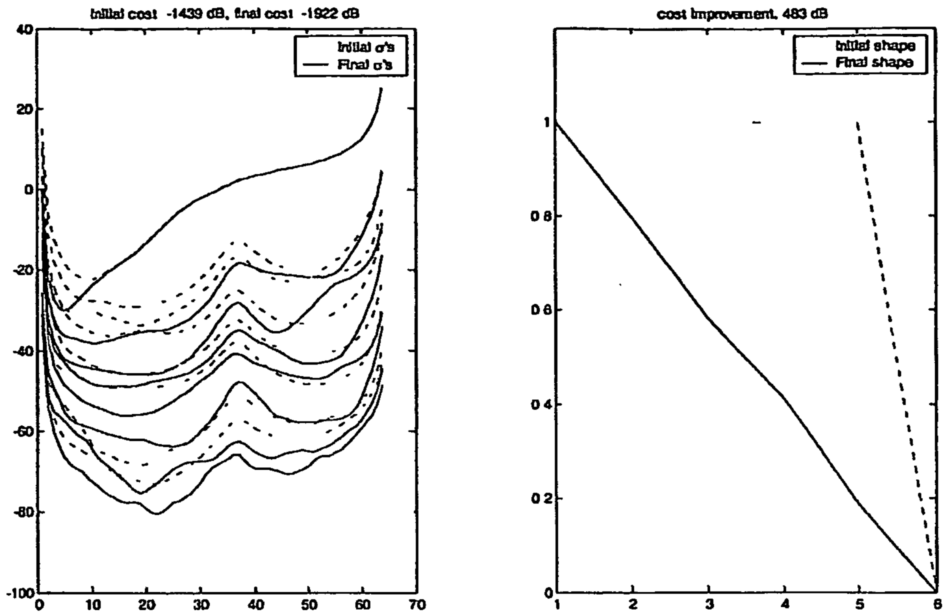
- window shape
- equalized cochannel set
- disturber symbol period (specified as a ratio of two integers)
- disturber symbol variance
- equalized main channel (for FEQ scaling purposes)
- length of the FFT in samples

The function produces the covariance matrices and their SVD's.

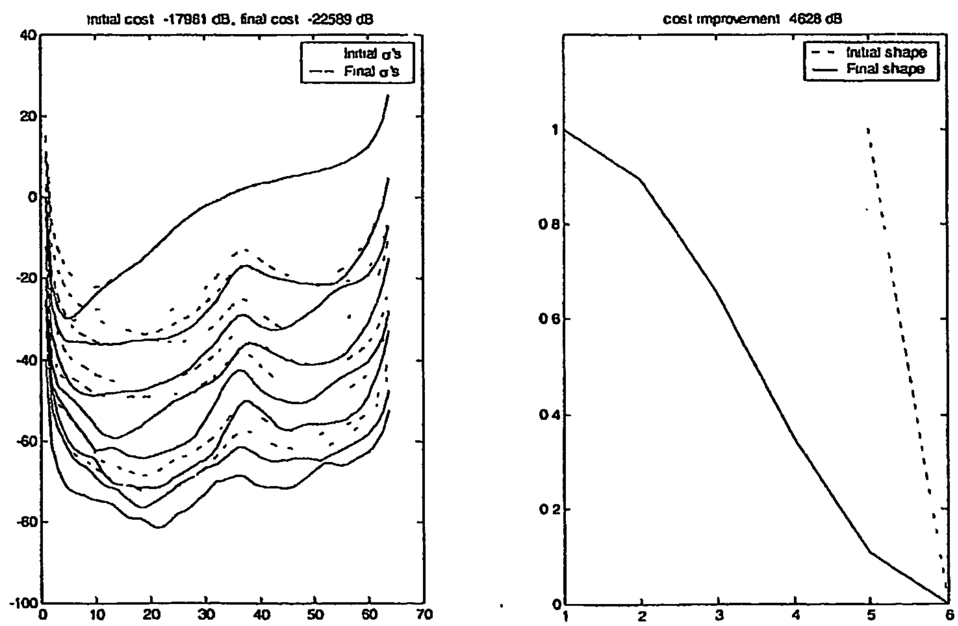
We then constructed a cost function wrapper that computes the double sum in expression (2). We then optimized this cost function with the MATLAB function `fminsearch()`, which implements a Nelder-Mead unconstrained optimization algorithm. We ran the optimization code with data from the Simulink time simulation. The results are shown in the sections below.

We also optimized just the sums over all bins of the first two singular values. This optimization makes sense if the smaller singular values are buried under white noise.

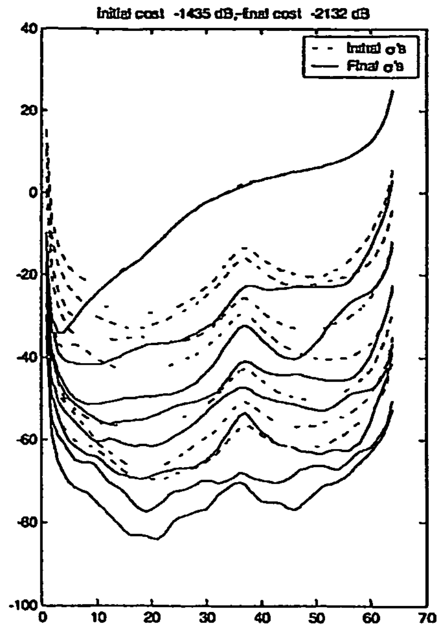
4.1 64-bins, 4-sample window, HDSL disturber, optimize σ_1 and σ_2

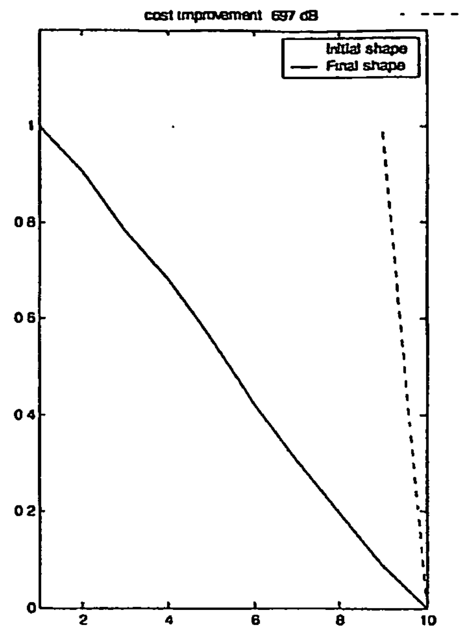


4.2 64-bins, 4-sample window, HDSL disturber, optimize all σ



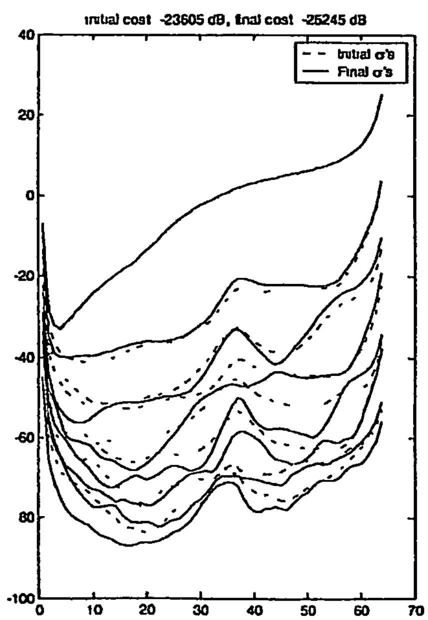
4.3 64-bins, 8-sample window, HDSL disturber, optimize σ_1 and σ_2

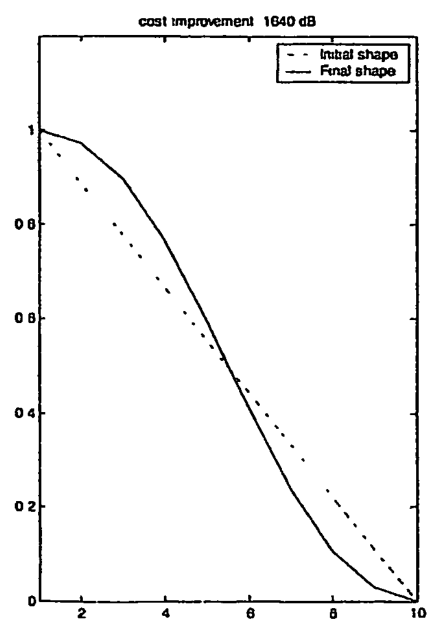




4.4 64-bins, 8-sample window, HDSL disturber, optimize all σ

(started from a linear window to speed convergence)





Subspace TEQ Design for DMT Systems

1. Introduction

Time domain equalization technique is used ubiquitously in current DMT systems. The purpose of the time domain equalizer (TEQ) is to shorten the otherwise much longer physical channel to within the prefix length such that the intersymbol interference (ISI) can be eliminated. The common approach to equalization is the minimum mean square error (MMSE) criterion which, however, is only optimal for single carrier systems, because in a multi-carrier system minimizing the overall mean square error does not necessarily lead to the ultimate goal – maximum data rate. In effect, the performance of MMSE TEQ is far from satisfactory (see [1]).

Different optimization criteria have been explored to improve TEQ performance, among which are the minimum ISI (MinISI) solution, maximum shortening SNR (MSSNR) method, maximum geometrical SNR (MGSNR) approach, etc. Compared with the MMSE result, these equalizers demonstrate the capability of increasing the overall transmission data rate. Unfortunately, none of these methods is optimal: certain approximations are used in each criterion to simplify the optimization procedure; even the MGSNR approach, although optimal in theory, is suboptimal in implementation as in practice it is too complicated to realize a true MGSNR TEQ. The comparison between these time domain solutions is provided in [1], which shows that both MMSE and MGSNR TEQs introduce multiple notches in the equalized channel, resulting in a great loss of bit carrying capacity at the corresponding frequency bins, while MinISI and MSSNR equalized channel eliminates those nulls and thus achieves a higher data rate.

However, the performance of MinISI and MSSNR TEQs is somehow channel dependent. In our system, these TEQs do reduce the number of multiple notches appearing in MMSE and MGSNR TEQ design, but they do not eliminate all of them. Obviously, SNR in the neighborhood of those frequency bins with channel notches is comparatively lower and thus the overall data rate is decreased. Therefore, how to design a high performance TEQ remains a challenge. Notice that a frequency domain approach, called per tone equalization (PTEQ) [2], leads to a very robust and effective equalizer design. Nevertheless, PTEQ employs a completely different structure which may not be practical for the existing systems. This proposal is restricted to the time domain equalizers only.

In this document, we introduce a novel and effective TEQ design based on the MSSNR criterion combined with the subspace method. The resulting TEQ, which we call subspace TEQ, shows comparable performance to that of the optimal PTEQ.

2. Subspace TEQ

2.1 The Problem

Examining various TEQ designs at hand, we observed that both MinISI and MSSNR TEQs outperform other approaches due to their smaller number of notches present in the equalized channel and hence a higher bit rate. However, these two TEQs do not eliminate all the nulls; there are still a few of them spaced almost at equal distances in the frequency domain. These nulls at lower

frequencies where bits are usually heavily loaded are the culprits impairing the system performance. This observation motivated our approach proposed in this document.

Our purpose is to remove the notches in the equalized channel. There are different ways to tackle this problem. To directly design a TEQ with some new criterion to achieve this goal is difficult if not impossible. As the MSSNR/MinISI TEQ produces only a few nulls, one may ask: is it possible to smooth the equalized channel by further implementing one additional step? The answer to this question leads to the emergence of the two-step subspace TEQ.

2.2 MSSNR Criterion

Before we develop the idea of the subspace TEQ, we briefly introduce the MSSNR design on which the new method is based. Notice that MSSNR TEQ and MinISI TEQ have very similar performance. Since the former uses the channel information only and is less expensive to implement, we choose the MMSNR criterion in our first step TEQ design.

The MSSNR TEQ design intends to maximize the energy of the portion of the equalized channel impulse response which lies inside a window of length $\nu+1$ taps corresponding to the desired length of the shortened channel. At the same time the energy outside this window is constrained. Let ν be the prefix length and $\mathbf{w}$ be the TEQ response. Given channel convolution matrix $\mathbf{H}$, define $\mathbf{H}_{win}$ as the portion inside a window and $\mathbf{H}_{wout}$ as the section outside of a window. Then the inside-window energy can be expressed as

$$\mathbf{h}_{win}^H \mathbf{h}_{win} = \mathbf{w}^H \mathbf{H}_{win}^H \mathbf{H}_{win} \mathbf{w} = \mathbf{w}^H \mathbf{A} \mathbf{w},$$

where $\mathbf{A} = \mathbf{H}_{win}^H \mathbf{H}_{win}$, and the residual is

$$\mathbf{h}_{wout}^H \mathbf{h}_{wout} = \mathbf{w}^H \mathbf{H}_{wout}^H \mathbf{H}_{wout} \mathbf{w} = \mathbf{w}^H \mathbf{B} \mathbf{w},$$

where $\mathbf{B} = \mathbf{H}_{wout}^H \mathbf{H}_{wout}$. Thus, the MSSNR criterion can be formulated as

$$\max_{\mathbf{w}} \mathbf{w}^H \mathbf{A} \mathbf{w} \qquad \text{subject to } \mathbf{w}^H \mathbf{B} \mathbf{w} = 1.$$

Obviously, this is a maximum eigenvalue problem and the solution can be found as

$$\mathbf{w}_{opt} = (\sqrt{\mathbf{B}}^H)^{-1} \mathbf{l}_{\max}$$

where $\mathbf{l}_{\max}$ is the unit length eigenvector associated with the maximum eigenvalue $\lambda_{\max}$ of the composite matrix $\mathbf{C}$ which is defined as

$$\mathbf{C} = (\sqrt{\mathbf{B}})^{-1} \mathbf{A} (\sqrt{\mathbf{B}}^H)^{-1}.$$

Thus, the MSSNR procedure searches the maximum eigenvalue of $\mathbf{C}$ for different delay and the MSSNR TEQ corresponds to the one that has the largest eigenvalue among all possible delays.

Generally matrix $\mathbf{B}$ is positive definite and hence $\sqrt{\mathbf{B}}$ can be computed via the Cholesky decomposition. However, $\mathbf{B}$ is channel dependent and in some cases it does not satisfy the positive definiteness requirement due to numerical inaccuracies. A simple and effective way to fix this problem is to pre-condition the matrix by adding a small positive term, which is based on the fact that if $\mathbf{B}$ is not positive definite, its negative eigenvalue is usually very close to zero, and thus a

small conditioning number will sufficiently remove the negative eigenvalues from $\mathbf{B}$. This modification leads to $\overline{\mathbf{B}} = \mathbf{B} + \beta\mathbf{I}$, where $\beta > 0$ assumes a comparatively small value and $\overline{\mathbf{B}}$ will be used instead of $\mathbf{B}$ in the above equations.

2.3 Subspace TEQ

The rationale behind the subspace TEQ method originates from the following observation: the maximum eigenvalue of matrix $\mathbf{C}$ associated with the optimal delay sometimes is not dominant, which means that there exist a few eigenvalues, though smaller than the largest one, still comparable to the largest one and much larger than the rest. Then the question is: can we combine these eigenvalues instead of using the largest one only to find a better TEQ?

Our subspace TEQ explores the above idea to combines the subspace method with the MSSNR design rule. The resulting TEQ performs a two-step optimization procedure: first, the optimal delay is determined using the MSSNR criterion and a few dominant eigenvectors of the composite matrix corresponding to this delay are obtained. Second, optimization is performed over the subspace spanned by these eigenvectors and the optimal TEQ coefficients are computed.

Let $\lambda_{\max,1} > \Lambda > \lambda_{\max,K}$ be the K largest eigenvalues of $\mathbf{C}_{opt}$ which is the composite matrix computed at the optimal delay. The corresponding eigenvectors are denoted as $\mathbf{l}_{\max,1}\ \Lambda\ \mathbf{l}_{\max,K}$. Rather than using $\mathbf{l}_{\max} = \mathbf{l}_{\max,1}$ in determining the optimal TEQ (in the sense of maximum shortening SNR) as $\mathbf{w}_{opt} = (\sqrt{\mathbf{B}}^{H})^{-1}\mathbf{l}_{\max}$, we may choose to use $\mathbf{l}_{\max} = \alpha_1\mathbf{l}_{\max,1} + \Lambda + \alpha_K\mathbf{l}_{\max,K}$ instead, where $\alpha_1, \Lambda, \alpha_K$ are positive scalars. We now have

$$\mathbf{w} = \alpha_1\mathbf{l}_{\max,1} + \Lambda + \alpha_K\mathbf{l}_{\max,K} = \mathbf{L}\boldsymbol{\alpha}$$

where

$$\mathbf{L} = [\mathbf{l}_{\max,1}\ \Lambda\ \mathbf{l}_{\max,K}], \qquad \boldsymbol{\alpha} = [\alpha_1\ \Lambda\ \alpha_K]^T.$$

So the next step is to optimize $\mathbf{w}$ by choosing a "good" $\boldsymbol{\alpha}$. As the ultimate goal of our system design is to achieve a higher bit rate, we choose to maximize the data rate over $\boldsymbol{\alpha}$ to obtain the optimal equalizer coefficients:

$$\mathbf{w}_{opt} = \mathbf{L}\boldsymbol{\alpha}_{opt}, \qquad \boldsymbol{\alpha}_{opt} = \arg\max_{\boldsymbol{\alpha}} f(\boldsymbol{\alpha})$$

where $f(\boldsymbol{\alpha})$ is a function of $\boldsymbol{\alpha}$ calculating the overall bit rate. Function $f(\boldsymbol{\alpha})$ is calculated in two steps. First the SNR in each bin is calculated as a function of the TEQ impulse response (and hence as a function of $\boldsymbol{\alpha}$). Then, the bit-loading formula is applied to the calculated SNRs.

This cost function is a non-linear one and requires gradient-based optimization procedures. Other simpler suboptimal cost functions are clearly also possible. The proposed framework allows a linear parameterization of the TEQ impulse response within a subspace that guarantees good ISI properties and facilitates further optimization within that subspace.

3. Performance Results

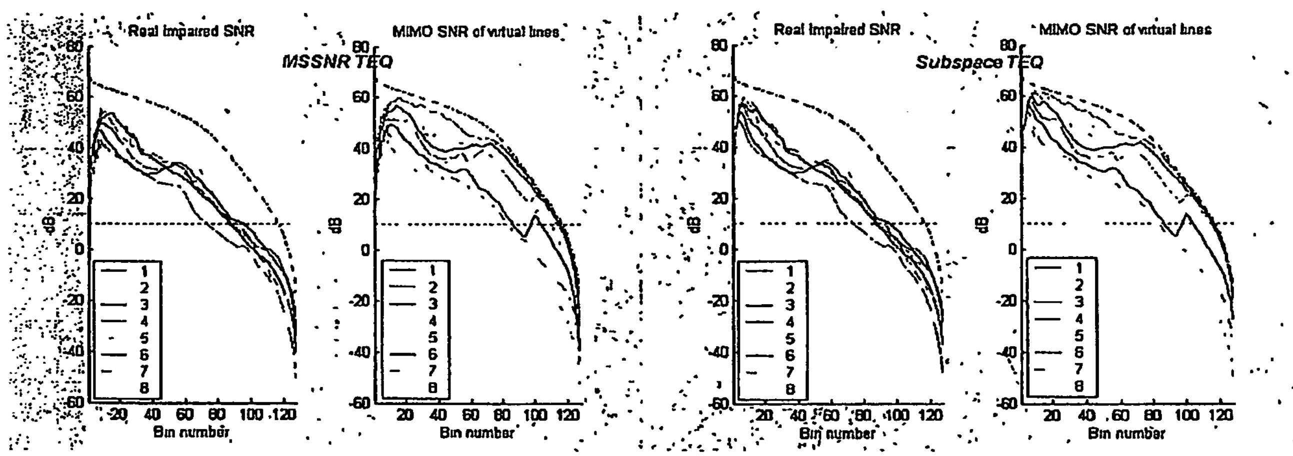
In this section, we provide some test results to demonstrate the performance of the proposed subspace TEQ.

Simulation performed here is based on a practical system setting. We consider a 25-pair binder with 8 lines deployed at 9Kft. The impulse responses of the channels are generated using LineMod. Crosstalk models come from lab measurement. Two disturber scenarios are considered next: (1) a moderately impaired case and (2) a heavily impaired case. From the following table, it is shown that compared with the MSSNR TEQ, the subspace TEQ increase the data rate by 1Mbps in both scenarios. What is well worth notice is that the performance of the subspace TEQ is comparable to that of the frequency per tone equalizer.

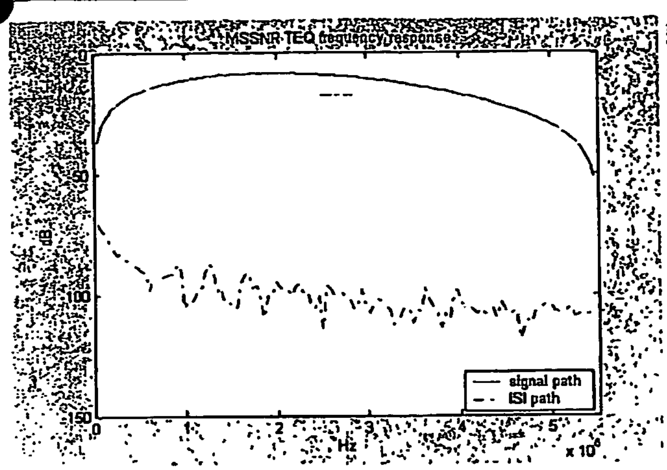
Data rate (Mbps)	No TEQ	MSSNR TEQ	Subspace TEQ	PTEQ
Moderately impaired	39.44	37.18	38.18	38.07
Heavily impaired	32.86	30.73	31.79	31.75

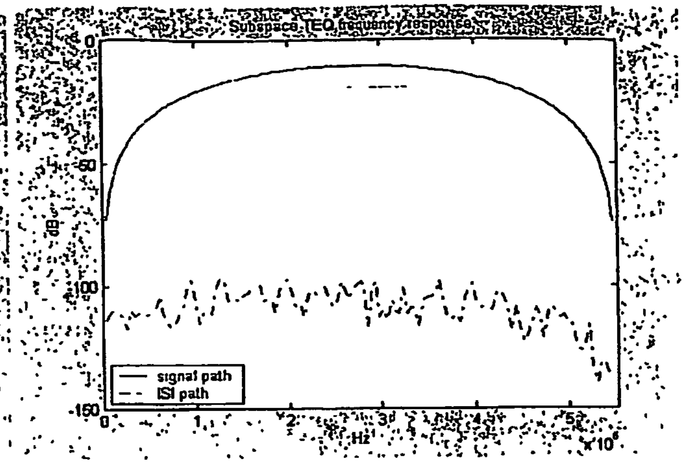
Table 1: Bit rate in Mbps without TEQ, with MSSNR TEQ and with subspace TEQ.

The difference can be explained by the receiver SNR plotted below. The 1Mbps loss of MSSNR TEQ happens at low frequencies where the subspace TEQ pushes the SNR up a few dBs. Because this region is heavily loaded, a little gain in SNR buys us one additional Mbps in bit rate.

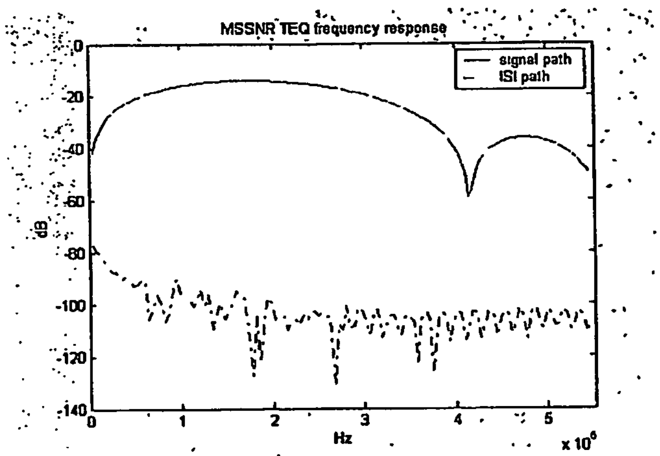


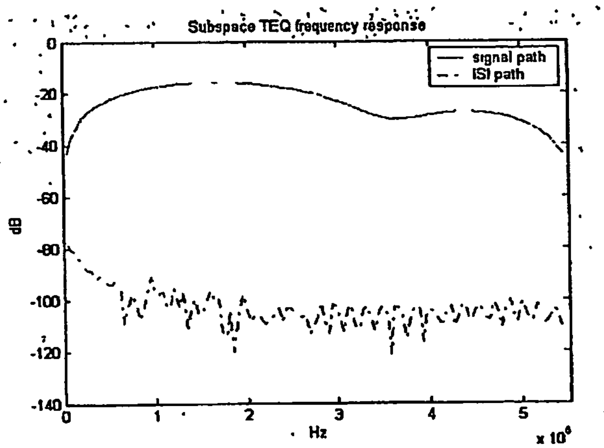
In this case, the MSSNR TEQ does not produce a notched equalized channel, nor does the subspace TEQ. The difference between two equalizers lies in the fact that the signal-to-ISI ratio of the combined channel and TEQ response is relatively higher at low frequencies if the subspace TEQ is used.





To see how the subspace smoothes the frequency notches, we examine another example where bridge-tapped loops are deployed. The system adopting the subspace TEQ gains about 0.5Mbps over the one with MSSNR TEQ. The frequency response of the equalized channel is presented below. It is clear that the subspace TEQ almost removes the null at about 410KHz. Because the overall SNR in this region is not high due to channel attenuation, the resulting data increase of the subspace TEQ is not substantial.





Multi-Pair Optimization

Introduction

Consider the problem of transport at very high bit rate over multiple copper pairs with a two-sided transceiver solution. Applications include symmetric (T3 / fractional services) and asymmetric (10 Base T/ 100 Base T Transparent LAN Service) data rates up to 10 Mbit/sec per pair at distances up to 9kft on 26 guage wire.

Spectral Management Standard Issue-1 defines two methods of establishing compatibility at the CO. Issue-1 does not address compatibility at a Remote Site or an Intermediate Point. Of these two methods, the first one uses transmission psd masks and templates for services at specific loop lengths (Equivalent Working Length). The second method uses an analytical approach to computing crosstalk from a proposed new technology service in to existing services (SS: Reverify) as well as from existing services in to the proposed new technology service. The NRIC enforcement mechanism is still based on having the affected party providing evidence of spectral incompatibility (SS: Reverify)

This document provides some technical details relating to earlier disclosures [ ]of spectral optimization based on allowable budget of ingress into specific service types that exist on each pair in the effective binder.

Problem Formulation

Notation:

$X_{In}(\omega)$: Vector of Transmitted signals from In-domain Transmitters.

$X_{Out}(\omega)$: Vector of Transmitted signals from Out of domain Transmitters.

$Y_{In}(\omega)$: Vector of Received signals at In-domain Receivers.

$Y_{Out}(\omega)$: Vector of Received signals at Out of domain Receivers.

$$X(\omega)=\begin{bmatrix} X_{In}(\omega) \\ X_{Out}(\omega) \end{bmatrix}, \; Y(\omega)=\begin{bmatrix} Y_{In}(\omega) \\ Y_{Out}(\omega) \end{bmatrix}$$

$$X_{In}(\omega)=\begin{bmatrix} X_{In-CO}(\omega) \\ X_{In-CPE}(\omega) \end{bmatrix}, X_{Out}(\omega)=\begin{bmatrix} X_{Out-CO}(\omega) \\ X_{Out-CPE}(\omega) \end{bmatrix}$$

$$Y_{In}(\omega)=\begin{bmatrix} Y_{In-CO}(\omega) \\ Y_{In-CPE}(\omega) \end{bmatrix}, Y_{Out}(\omega)=\begin{bmatrix} Y_{Out-CO}(\omega) \\ Y_{Out-CPE}(\omega) \end{bmatrix}$$

$$Y(\omega)=\begin{bmatrix} Y_{In}(\omega) \\ Y_{Out}(\omega) \end{bmatrix}=\begin{bmatrix} H_1(\omega) & H_2(\omega) \\ H_3(\omega) & H_4(\omega) \end{bmatrix}\begin{bmatrix} X_{In}(\omega) \\ X_{Out}(\omega) \end{bmatrix}+\nu$$

$\Phi_{XX}(\omega)$ is the Spectral Density of X and similarly for other signals.

Various Spectral Optimization problems can now be posed.

Problem 1 (Symmetric Service with Ingress Budget):

Problem 1.a

Deliver the highest data rate multi-pair symmetric service at a given error rate under the constraint that ingress from it into other services in the binder is no worse than that allowed under the current Spectral Management Standard.

Unlike the convention that the same transmit spectrum be used for the proposed service for all possible out-of-domain service deployment scenarios at a particularly loop length, this formulation allows transmit spectrum dependent on the actual deployment scenario. Moreover, the transmit spectrum optimization utilizes knowledge of cross-talk transfer functions. The issues of uncertainty in cross-talk transfer functions and the need for real-time adaptation will be addressed later.

Given a specific out-of-domain disturber deployment with the corresponding $X_{Out}(\omega)$ and given all relevant transfer functions, we pose the following mathematical optimization problem.

$$\boxed{\max_{\Phi_{XX_{in}}} r = r(\Phi_{XX_{in}}, H, \Phi_{XX_{out}}, R_v) = rate_{upstream} + rate_{downstream}}$$

Subject to:

$$\boxed{\begin{array}{l} rate_{upstream} = rate_{downstream}, \\ \Phi_{YY_{Out\,ii}}(\omega) \le \Delta_{SMbound_i}(\omega), \qquad i = 1, \Lambda\ N_{Out} \end{array}}$$

The $rate_{upstream}$ and $rate_{downstream}$ functions are dependent on the chosen transmitter / receiver architecture. For example, for MMSE treatment of out-of-domain receiver in OFDM transmitter / receiver architecture,

$$rate_{downstream} = \sum_{i=1}^{Nbins} \log_2 \det(I + \overline{H}\Phi_{XX_{in-CPE}}\overline{H}^*)\ df$$

where

$$\begin{aligned} Y_{in-CO} &= H_{1CO-CO}X_{in-CO} + H_{1CO-CPE}X_{in-CPE} + H_2 X_{out} + v_{in-CO} \\ &= Y_{in-NEXT} + H_{1CO-CPE}X_{in-CPE} + Y_{dis} \\ R_{YY_{dis}}^{-1/2}(Y_{in\_CO} - Y_{in-NEXT}) &= R_{YY_{dis}}^{-1/2}H_{1CO-CPE}X_{in-CPE} + R_{YY_{dis}}^{-1/2}Y_{dis} \\ \overline{H} &= \Gamma^{-1/2}R_{YY_{dis}}^{-1/2}H_{1CO-CPE} \end{aligned}$$

Definition:

$\Delta_{SMbound_i}(\omega)$ is the worst case egress budget on the out of domain service type i at the specific EWL

$$\max_{\text{all allowable delpoyments of } \Phi_{XX} \text{ under SM Standard}} \Phi_{YY_{Out_i}}(\omega)$$ using the worst-case Unger cross-talk transfer function mask.

Note that at each frequency, the worst case egress may occur due to a different combination of allowable disturber deployment.

Problem I.b

An alternative approach to egress budget is to use the specific receiver margin calculation method in SM standard and then constrain the optimization to allow margin on the i^{th} out-of-domain service to be at that limit using knowledge of the actual transfer functions and the actual in-domain and other out-of-domain services.

ABSTRACT

A method and system are disclosed using multi-lines to deliver ultra high speeds in a communications system.